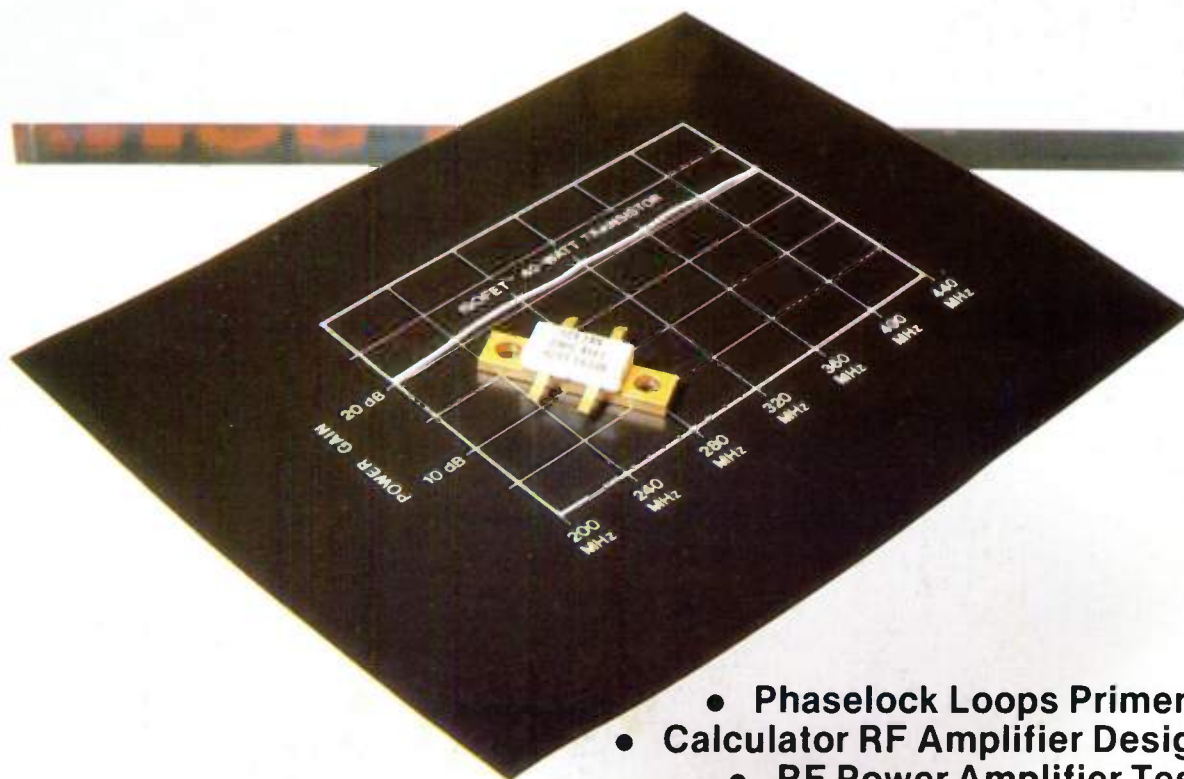


# rf design

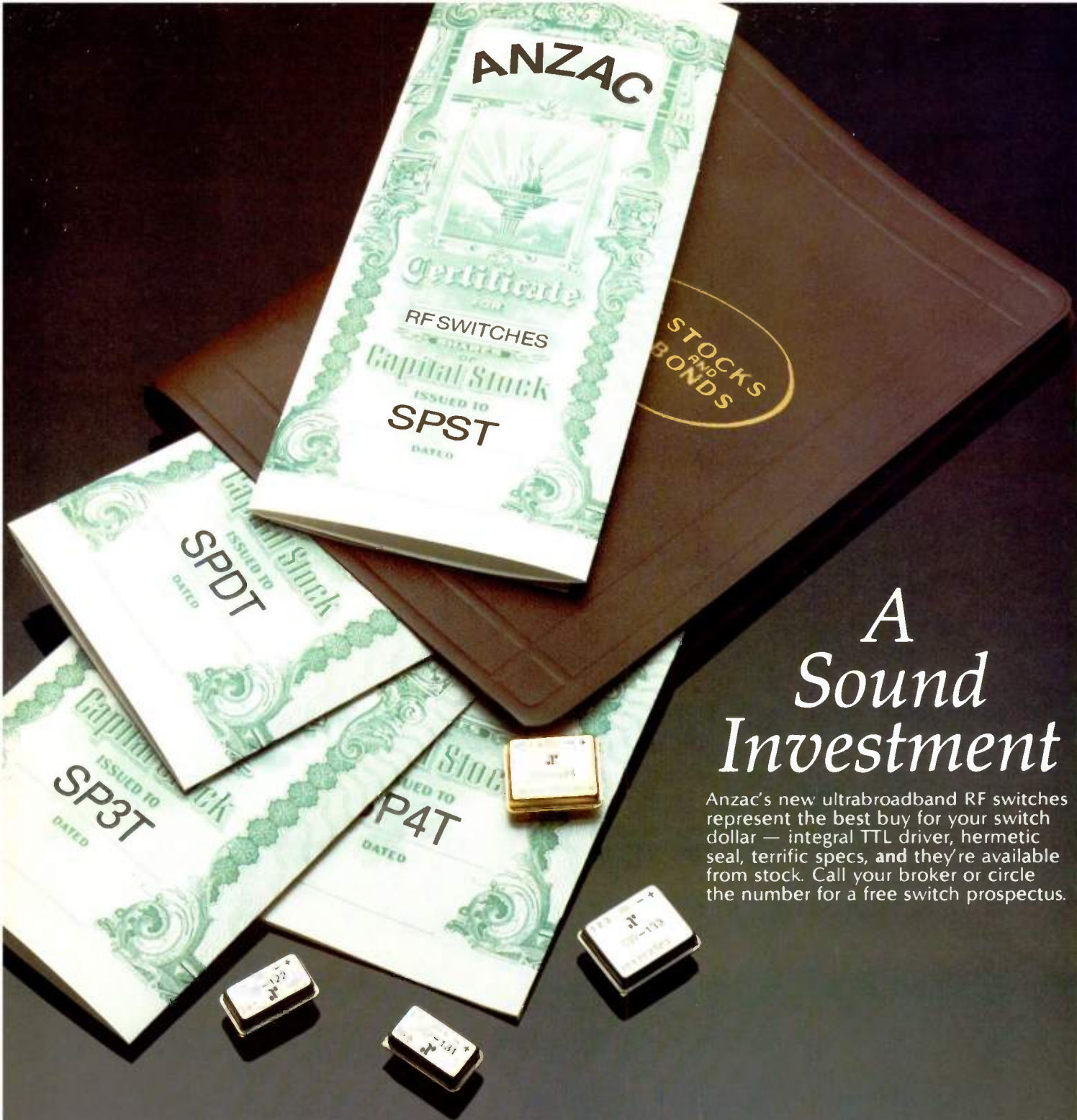
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Frequency Range (MHz)	5 to 2000	5 to 1000	200 to 2000	5 to 2000	5 to 1000	200 to 2000	5 to 2000	5 to 1000	200 to 2000	5 to 2000	5 to 1000	200 to 2000
VSWR	1.1	1.05	1.15	1.1	1.05	1.15	1.1	1.1	1.2	1.1	1.1	1.2
Isolation (dB)	80	60	50	80	60	50	60	60	45	60	60	45
Insertion Loss	1.2	0.5	0.5	1.2	0.5	0.5	1.3	0.6	0.6	1.3	0.6	0.6

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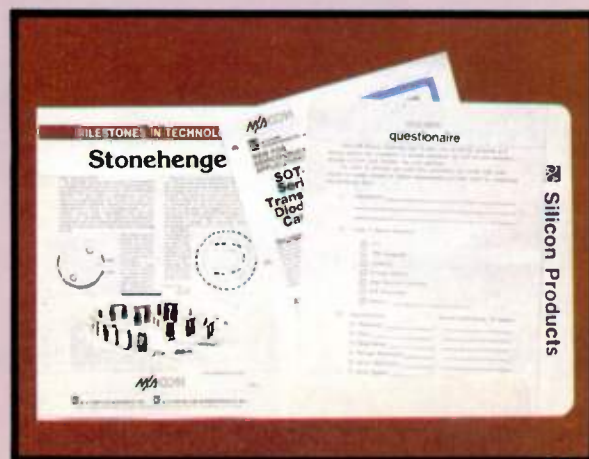


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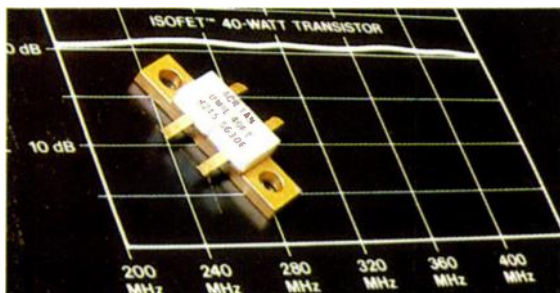
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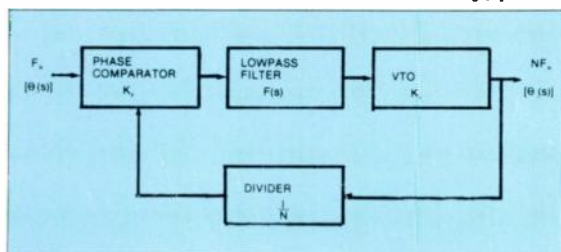
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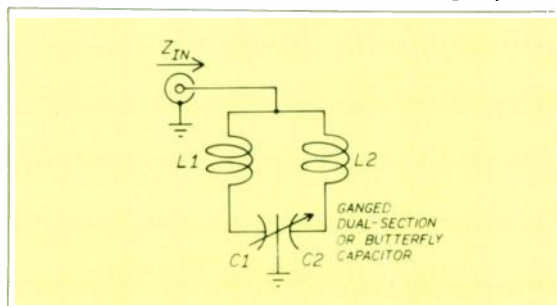
Cover Story, p. 70.



PLL Primer, p. 18.



RF Amplifier Design, p. 32.



RF Power Amp Testing, p. 48.

**March/April Cover.** The cover, courtesy of Acrian, Inc., announces a new MOSFET fabrication technique enabling power amplifier designers in the VHF and microwave regions to design amplifier stages with higher gain, better isolation and thermal stability. See the cover story on page 70.

**PLL Primer: Part I.** Taking the mystery out of phase-lock loop design — an integrated approach.

**RF Amplifier Design Program.** A calculator software package able to evaluate RF amplifier performance and determine required source and load terminations for any condition of stability.

**Lumped-Constant Line Stretcher For Testing Power Amplifier Stability.** To determine the stability of an RF power amplifier, oscillator, or transmitter under harsh conditions, variable line and fixed attenuator are used to simulate a given VSWR at all phase angles. This article describes a lumped-constant line stretcher using either variable capacitors or inductors.

**An Easy Method for Measuring Unloaded Q (Q<sub>u</sub>) for an Inductor or Filter Tank.** A measurement technique to measure inductance by using a narrow-band single pole bandpass filter whose Q approximately equals the unloaded Q of the inductor or tank being measured.

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Cover Story	70	New Literature	80
Feature Products	72	Ad Index	82



# pick a mixer



SRA-1



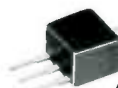
SBL-1X



SBL-1



TFM-2



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- SBL-1X industrial grade, low cost, \$4.95 (10-49) 10 to 1000 MHz, rugged all metal enclosure.
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\*Units are not QPL listed

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MODEL	SRA-1	TFM-2	SBL-1	SBL-1X	ASK-1
<b>FREQUENCY, MHz</b>					
LO, RF	5-500	1-1000	1-500	10-1000	1-600
IF	DC-500	DC-1000	DC-500	5-500	DC-600
<b>CONVERSION LOSS, dB</b>					
one octave bandedge	6.5	6.0	7.5	7.5	7.0
total range	8.5	7.0	8.5	9.0	8.5
<b>ISOLATION, dB, L TO R</b>					
lower bandedge	50	50	45	45	50
mid range	40	40	35	30	35
upper bandedge	30	30	25	20	20

For complete specifications and performance curves refer to the 1980-1981 Microwaves Product Data Directory, the Goldbook or EEM.

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C-74 REV C

# letters

Dear Sir:

After reading the article "Triple Tuned Circuits" by Andrzej B. Przepelski in your July/August issue of *r.f. design*, I as a reader was left with the opinion this method of design could be improved in certain areas.

In order to get exact values and symmetrical roll-offs (18dB/octave) on each side of the filter, the following method of design is suggested:

(A) Design Objectives

BW = 40 KHz

Rin = 25 ohms

Rout = 75 ohms

$f_1 = -3\text{dB Frequency Low Side}$

$f_2 = -3\text{dB Frequency High Side}$

$f_c = 36.64 \text{ KHz} = \sqrt{F_1 \times F_2}$

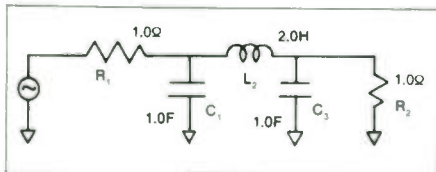
$n = 3$  (18dB/octave roll-off)

(B) Design a low pass filter, Butterworth type with  $n = 3$ , and a cut-off frequency equal to 40 KHz (desired bandwidth) with an input and output impedance equal to 25 ohms.

(C) Using the Bartlett Bisection Theorem and scale the load impedance and associated components to 75 ohms (see appendix).

(D) Complete the filter by resonating (at the center frequency) each shunt capacitor with a shunt inductance and a series capacitor with the series inductor.

Example: Look up the normalized values of a Butterworth Low Pass Filter in any of the following texts on filters: Zveref, Weinberg, White or Geffe. ( $n = 3$ )

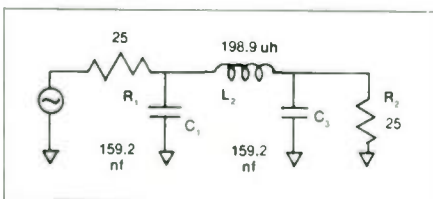


$$\omega_c = 2\pi F = 1 \text{ Radian}$$

Scale to 25 ohms ( $R_1$ ) and 25 ohms ( $R_2$ ).

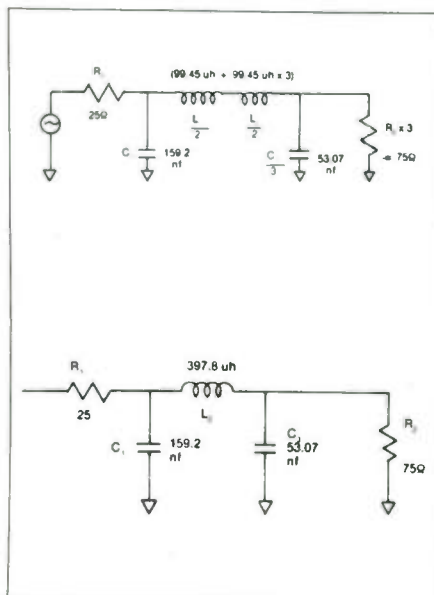
$$C_1 = C_3 = \frac{1}{25} \times \frac{1}{2\pi 40 E3} \times 1 = 159.2 \text{ nF}$$

$$L_2 = \frac{25}{1} \times \frac{1}{2\pi 40 E3} \times 2 = 198.9 \mu\text{H}$$



$$\omega_c = 2\pi 40 \text{ KHz} = 251.3 \times 10^3 \text{ Radians}$$

(E) Use Barlett Bisection Theorem convert load impedance to 75 ohms with associated modified elements.

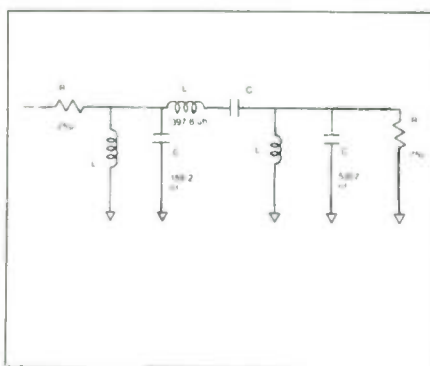


Low Pass Filter, Cutoff = 40 KHz

Rin = 25 ohms

Rout = 75 ohms

(F) Convert to Band Pass Filter



Band Pass Filter  $f_0 = 34.64 \text{ KHz}$

$$L_1 = \frac{1}{(2\pi 34.64 E3)^2 159.2 E-9} = 132.6 \mu\text{H}$$

$$L_3 = \frac{1}{(2\pi 34.64 E3)^2 53.07 E-9} = 397.8 \mu\text{H}$$

$$C_2 = \frac{1}{(2\pi 34.64 E3)^2 397.8 \mu\text{H}} = 53.07 \text{ nF}$$

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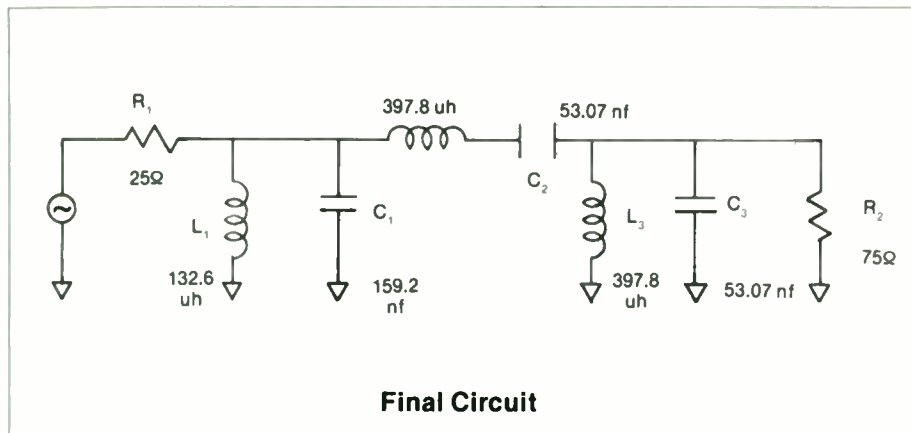
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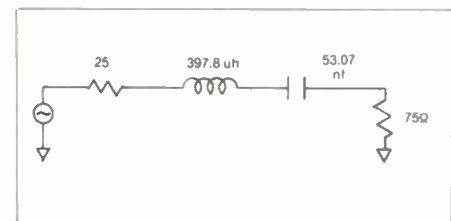
Band Pass Filter BW = 40 KHz  
 $f_0 = 34.36 \text{ KHz}$   
 Roll-Off =  $2\pi n/\text{octave}$  where  $n = 3$   
 = 18dB/octave

BW Check (parallel ckt.)

$$BW = \frac{1}{2\pi CR} = \frac{1}{2\pi 159.2E-9 \times 25} = 40 \text{ KHz}$$

$$BW = \frac{1}{2\pi 53.07E-9 \times 75} = 40 \text{ KHz}$$

At Resonance, the Zp of  $L_1$  and  $C_1$ , and  $L_3$  and  $C_3$ , in parallel, are very large and the circuit becomes



BW check (series circuit)



$$BW = \frac{R}{2\pi L} = \frac{100}{2\pi 397.8 \text{ uH}} = 40 \text{ KHz}$$

$$Q = \frac{1}{R_s} \sqrt{L/c}$$

$$Q = \frac{1}{100} \sqrt{\frac{397.8 \text{ uH}}{53.07 \text{ nF}}} = .8658$$

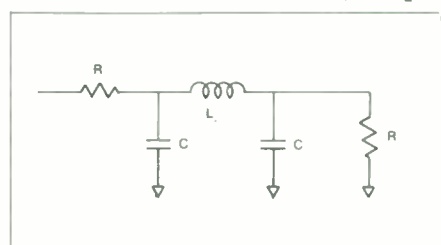
$$BW = \frac{f_0}{Q}$$

$$BW = \frac{34.64E3}{.8658} = 40 \text{ KHz}$$

## Appendix

Bartlett Bisection Theorem enables an RF designer to transform an equally terminated Low Pass Filter to a Low Pass Filter with unequal terminations.

Given Low Pass Filter with  $R_1 = R_2$



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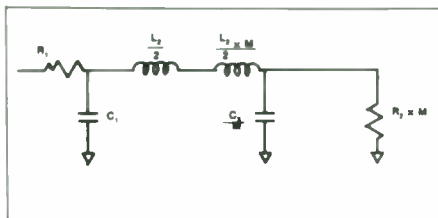
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DETERMINE: Output load resistance desired, to input resistance ratio, M

$$\frac{R_2}{R_1} = M \text{ where } R_1 < R_2$$

$$M < 10$$

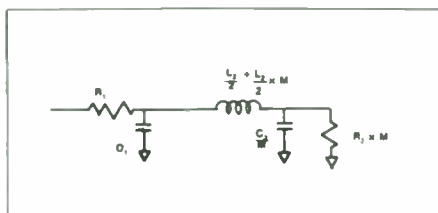


Divide  $L_2$  by two; multiply one-half by M and sum the results —

$$\left( \frac{L_2}{2} + \frac{L_2}{2} \times M \right)$$

Divide  $C_3$  by M  
Multiply  $R_2$  by M

### Final Filter



I enjoy reading *r.f. design* and find it a very interesting and informative magazine.

Tony Tortorella

Dear Editor: Tortorella is correct. His solution, even though not completely exact (it neglects the Q of the inductors), is a more elegant one for unequal terminations. The final roll-offs of both filters is 18dB/octave, since both are third order filters. The simple method described in the article was basically meant to be used for equal termination applications and the unequal termination example was used only to show that it can also be used in these cases with minimal sacrifice in performance.

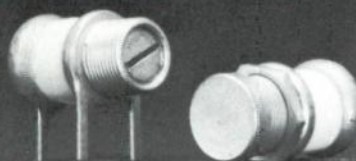
The enclosed figure shows the difference in response between the two filters. The main difference is in the high frequency performance. The equal termination design (when used with unequal source and load) starts rolling off a little slower and does not reach the 18 dB per octave slope until it reaches about 90 KHz.

Using equal terminations both methods give the same filter component values and performance.

A. Przedpelski

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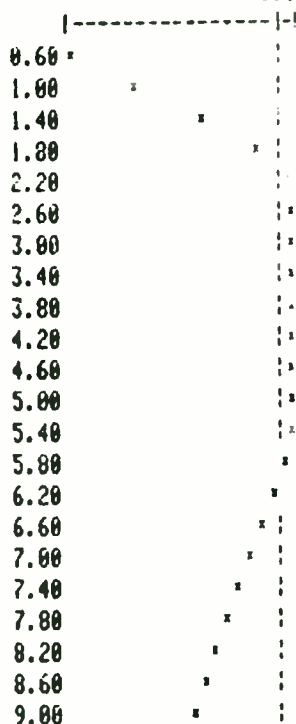
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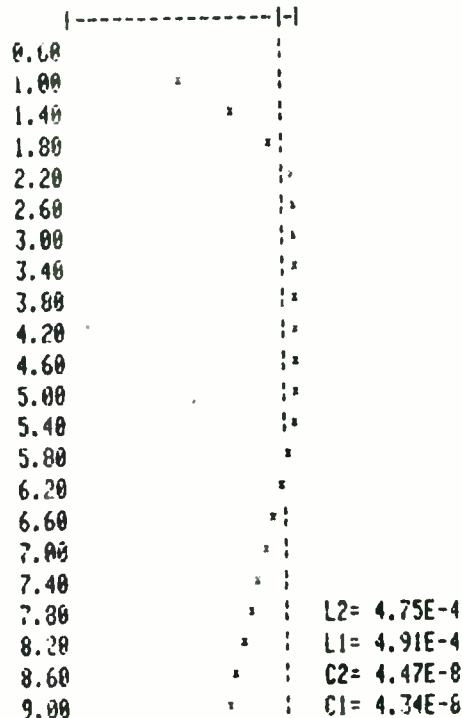
"Exact"  
solution

PLOT OF RESP  
X<UNITS = E4.>↓  
Y<UNITS = 1.>→  
-40.0 -3.0  
-6.0



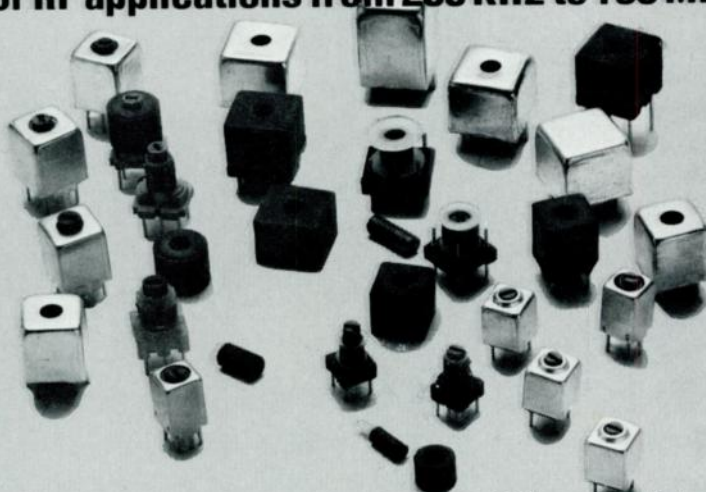
"Equal  
terminations"  
solution

PLOT OF RESP  
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Y<UNITS = 1.>→  
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-6.0

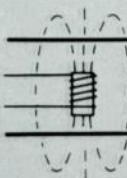


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Dear Sir:

The November/December 1982 issue of *r.f. design* contained an interesting and informative article by Andrzej B. Przedpelski entitled "Loop Antenna Design." There are several points not brought out in the article which may be of significance in low noise design. This letter discusses these points.

The loop equivalent circuit presented in the various figures contains a resistance  $R_L$ . It is never mentioned that this resistance is really the sum of a radiation resistance,  $R_r$ , and several components of loss resistance (total loss resistance =  $R_L$ ). An efficiency factor,  $n$ , can be defined for the antenna and is given by

$$n = \frac{R_r}{R_r + R_L} = \frac{R_r}{R_L}$$

For many loops, the loss resistance is much higher than the radiation resistance. The primary loss mechanisms are the ohmic loss in the wire, the losses introduced by the ferrite material, and losses caused by proximity effects when multiple turns are used. The efficiency impacts upon the overall system noise figure.

The article also does not mention the effects of atmosphere and man-made noise on received signal to noise ratio. Optimum performance for low noise applications at frequencies below 30 MHz occurs when the system is atmospherically noise limited. The noise figure of a lossless antenna can be defined as

$$f_a = \frac{\text{available noise power rec'd by antenna}}{kTB}$$

(this is, of course, akin to the concept of antenna temperature). The loss occurring in the antenna can be accounted for by introducing a lossy network at the output of the lossless portion of antenna. When the antenna is connected to a receiver, the equivalent circuit is as shown in Figure 1a. The noise figure for this circuit including atmospheric noise is

$$f = f_a - 1 + \frac{f_{rec}}{n}$$

where  $f_{rec}$  is the receiver noise figure. For optimum performance in low noise applications,

$$f_a - 1 > f_{rec}/n$$

When this occurs, no further improvement in signal to noise can happen since the system is atmospherically noise limited. It is quite easy, depending upon the electrical size of the antenna, the frequency, the type of active device, etc., for the receiver noise figure ( $f_{rec}/n$ ) to greatly exceed atmospheric noise. It is therefore important to consider atmospheric noise and antenna efficiency.

Finally, the article does not completely model the noise performance of the antenna/receiver combination. Figure 1b shows an equivalent noise circuit using one of the standard noise models for a two port network. This circuit corresponds to Figure 6 in the article.  $\bar{e}n^2$  and  $\bar{i}n^2$  are the equivalent noise sources of the amplifier referred to its input.  $Z_s$  is the impedance of the antenna/tuning circuit. The receiver's noise figure is

$$f_{rec} = 1 + \frac{\bar{e}n^2 + |Z_s|^2 \bar{i}n^2 + 2\text{Real}\{Z_s < \bar{e}n * \bar{i}n_s >\}}{4kTB \text{Real}(Z_s)}$$

where  $< \bar{e}n * \bar{i}n >$  is the correlation between  $\bar{e}n^2$  and  $\bar{i}n^2$ .  $\bar{e}n^2$  and  $\bar{i}n^2$  are often frequency dependent. Also depending upon the relative magnitude of  $\bar{e}n^2$  and  $\bar{i}n^2$ ,  $Z_s$ , which is also frequency dependent, can have a profound effect on  $f_{rec}$ . This is particularly true if  $\text{Imag}(Z_s) \gg \text{Real}(Z_s)$



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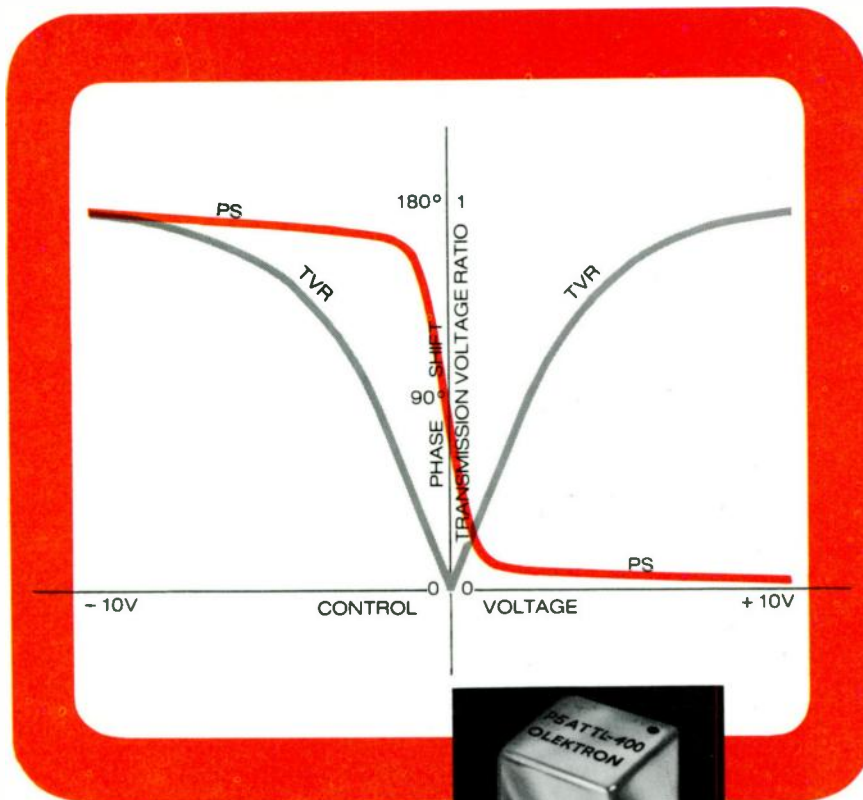
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MIL-C-83733. And, additional backshells are being qualified.

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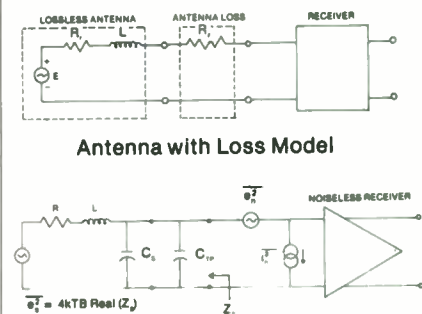
INFO/CARD 12

or if the real part of  $Z_s$  is small.

The points discussed above are of importance in low noise designs where optimum sensitivity is required. They are especially important in wide band designs where  $f_a$  can change by large amounts and the various frequency dependencies become significant.

Thank you for your consideration.

Sincerely,  
Robert Sainati



Antenna/Receiver Noise Circuit Model

**Figure 1. Antenna/Receiver  
Equivalent Circuits**

Dear Editor:

Following are comments on reader Sainati's letter:

Reader Sainati brings up interesting and theoretically valid points. Because of limited space, the article was meant to be a practical design aid since very little information is available in the literature.

1.  $R_L$  is defined in Fig. 2 as loss resistance due to coil Q only. Radiation resistance is neglected since as Snelling (Ref. 2) states, "it is usually negligible compared with the losses". This accounts for the well known low efficiency of common ferrite loop antennas (small effective height).

2. Only the method of optimizing the circuit was described, since atmospheric noise is not under the designer's control, and is usually not part of the receiver specifications. It is usually the "system" engineer who determines the equipment requirements and the "design" engineer who has to meet them.

3. The article stressed more the broadband rather than a single frequency design. Thus, the noise density rather than noise figure calculation was treated because of some of the problems described by reader Sainati. This approach has worked satisfactorily in broadband low frequency low noise applications in connection with long range propagation studies.

A. Przedpelski

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RF383

# PLL Primer

Taking the mystery out of phaselock loop design —  
an integrated approach.

By Andrzej B. Przepelski  
A.R.F. Products, Inc.  
R & D Laboratory  
Boulder, Colorado

**P**hase lock loops have been with us for decades<sup>1</sup> and yet there seems to be an aura of mystery associated with their design. Actually, their design is fairly straightforward, providing all factors are considered and the design procedure is closely followed. The majority of conventional PLLs (Figure 1) perform one of three functions:

- if  $N$  is unity, the PLL is a tracking filter;
- if  $N$  is larger than unity, but fixed, the PLL is a frequency multiplier;
- if  $N$  is a selectable value, the PLL is a frequency synthesizer.

While the above may be an oversimplification, it is adequate to illustrate the majority of PLL applications.

## Type and Order

The first step in PLL design is to select suitable PLL order and type. These are adequately described in a Motorola data book<sup>2</sup>. Briefly, in a conventional PLL, the order and type are determined by the lowpass filter configuration. Two types are in general usage: type 1, using a passive lowpass filter and type 2, using an active filter. Type 2 is usually the more desirable.

Second order PLL is the most commonly described in the literature for the above applications. Even Gardner<sup>3</sup> states, "It is certainly the most prevalent . . ." " . . . second-order loop, as commonly built, is unconditionally stable." However, he also adds: "Parasitic circuit elements often cause a loop intended to be a second order to actually be a much higher order." It is the author's contention that all second order loops are actually of a higher order, as many designers have found out when their "second order" loops oscillated. There always is a pole or two (or more) hidden

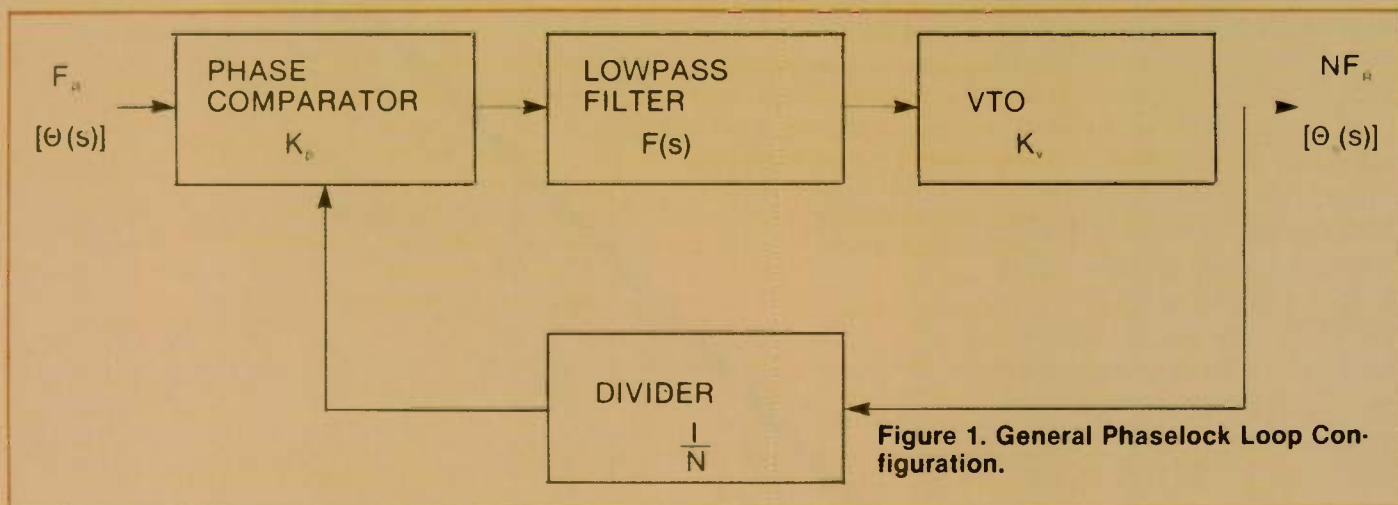
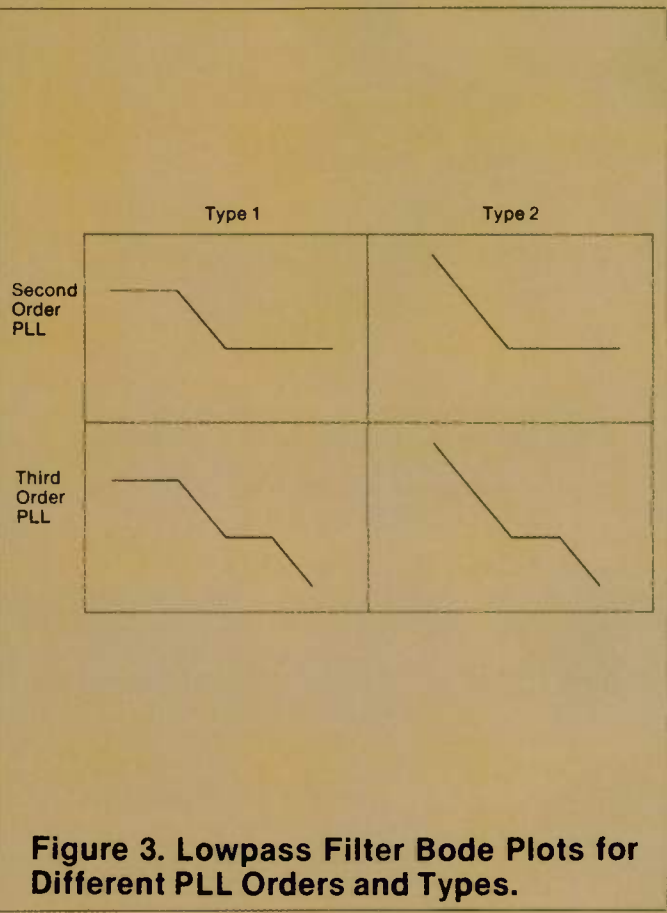
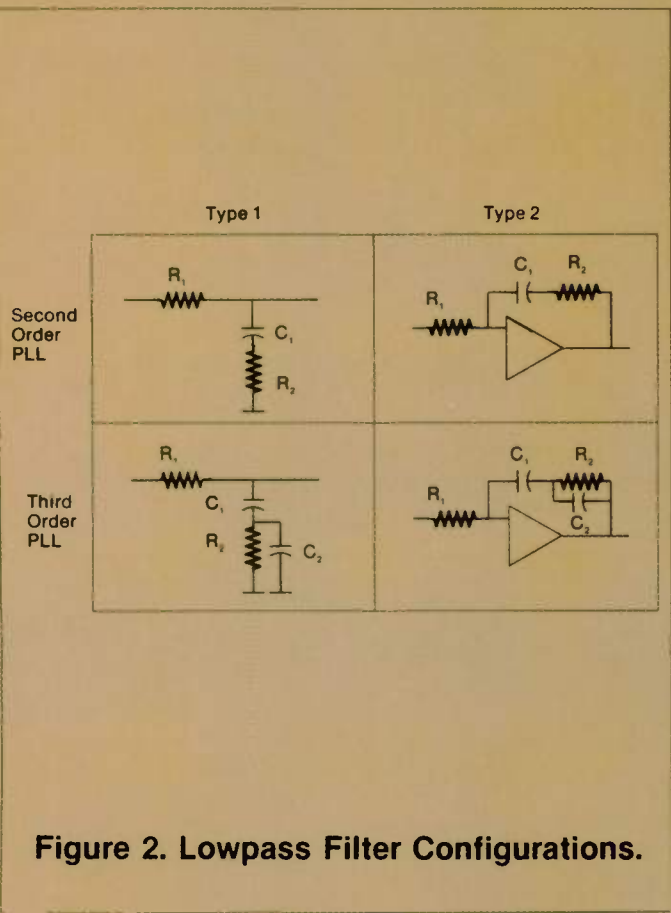


Figure 1. General Phaselock Loop Configuration.



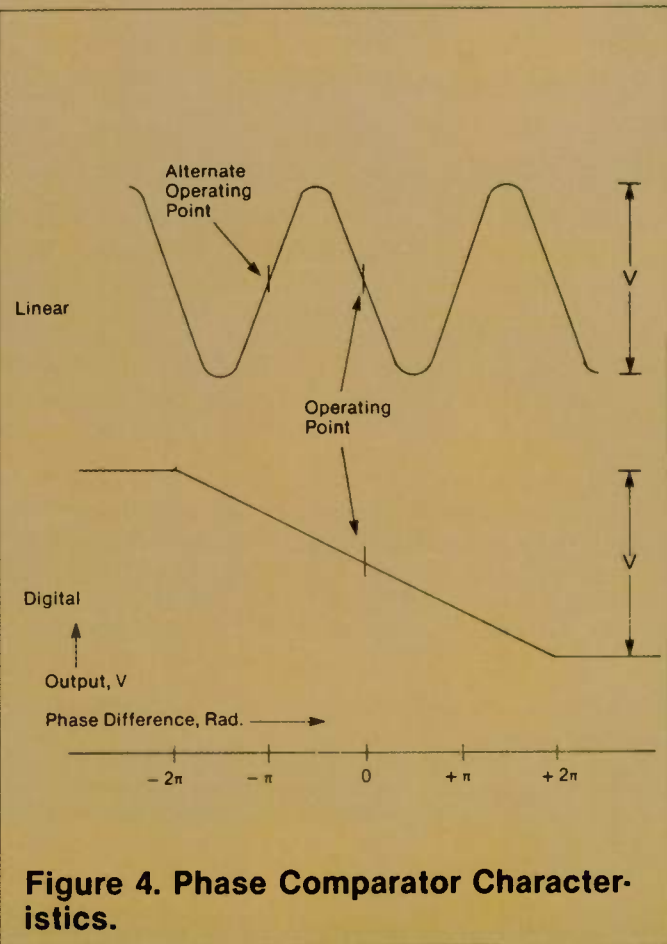
somewhere. Thus, it is more desirable (and less frustrating in the long run) to design a basically third order loop from scratch, especially since it has advantages over the second order. One important advantage is better suppression of reference frequency sidebands when  $N$  is more than one.

The lowpass filter configurations for the different PLL types and orders and their Bode plots are shown in Figures 2 and 3. Because of their inherent advantages, except in some special cases, only third order type 2 PLLs will be considered for further discussion, especially since the first and second orders are well covered in the references. Rohde<sup>4</sup> recognizes the validity of this approach and gives some examples.

## Phase Comparator

The next decision is the selection of phase detector type: linear or digital. The most common linear phase comparator is the double-balanced mixer. Except for special applications, such as low noise tracking filters<sup>5</sup>, the linear phase comparator is not the best choice. At first glance it seems that it has the advantage of providing a phase error output throughout the signal cycle. However, when  $N = 1$  the divided VTO signal contains phase information in zero crossings only. If the usual programmable divider type is used, this information is only in the leading edge. Another disadvantage is that the gain,  $K_p$ , varies with phase error. When no phase error exists (normal steady state operating condition) the gain is maximum and is  $V/2$  volts/radian (Figure 4). It then decreases to zero, when the loop starts skipping cycles. A small advantage of the linear phase comparator is that there is no need to worry about feedback polarity. The loop will choose an operating point where  $K_p$  polarity will provide overall negative feedback.

The usual digital phase comparator is actually a frequency/phase comparator, as shown in Figure 4. When





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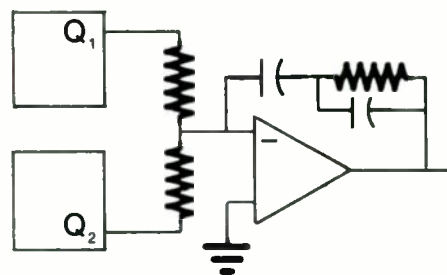
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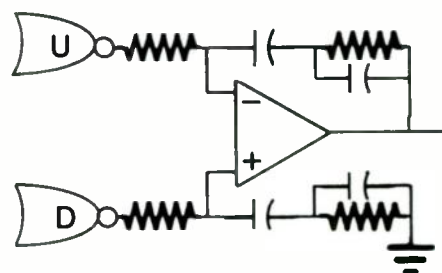


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**Figure 5. Digital Comparator/Integrator Interface.**

the phase error is larger than  $2\pi$  it provides a maximum output voltage with a polarity indicating direction of frequency error. This desirable feature, however, makes it necessary to determine the proper  $K_p$ ,  $K_v$  and  $F(s)$  polarities to provide an overall negative feedback in the loop. This is no problem if it is done before the PC board is made, since all that is required is to exchange the two comparator inputs, if the original polarity is not correct. The gain,  $K_p$ , of the conventional digital comparator is  $V/4\pi$  volts/radian. It should be pointed out that, while the curve of Figure 4 is commonly shown as the transfer curve, it is an "effective" and not actual output. The output is actually a series of pulses with a duty cycle which is a function of phase error. Several circuit configurations can be used. If a complementary symmetrical logic is used, such as CMOS, a differential configuration is suitable<sup>8</sup> and the two signals are summed in the integrator (lowpass filter) input; otherwise, the two signals are subtracted in a differential integrator<sup>2</sup>. These two configurations are shown in Figure 5. From these it can be seen that a digital frequency/phase comparator has two outputs which must be either added or subtracted. The digital version's main advantages are constant gain, large pull-in range and fast lock.

A slight glitch may be present in the TTL phase comparator characteristic when the on-chip charge pump is used. This may be undesirable in critical applications. The use of an external integrator reduces this problem considerably and using a faster logic comparator (with the external integrator) practically eliminates it<sup>9</sup>.

## Integrator

As discussed before, a type 2 PLL uses an active low-pass filter with a pole at zero frequency (ideally) or an integrator. The use of the integrator provides for much better

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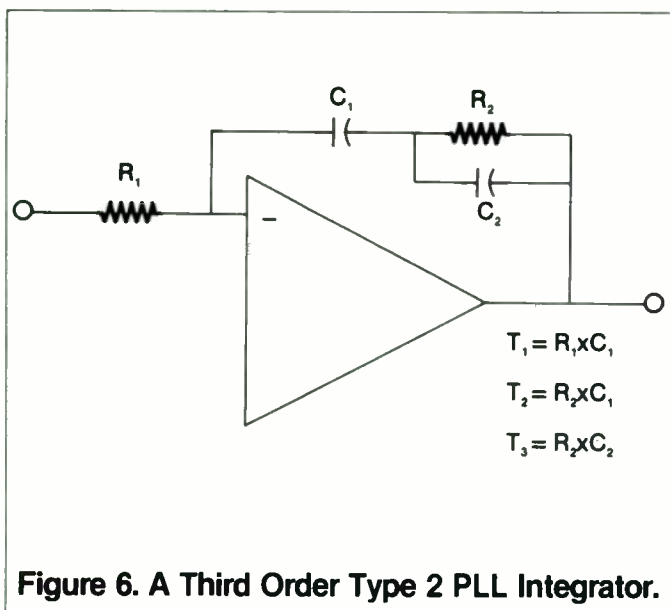
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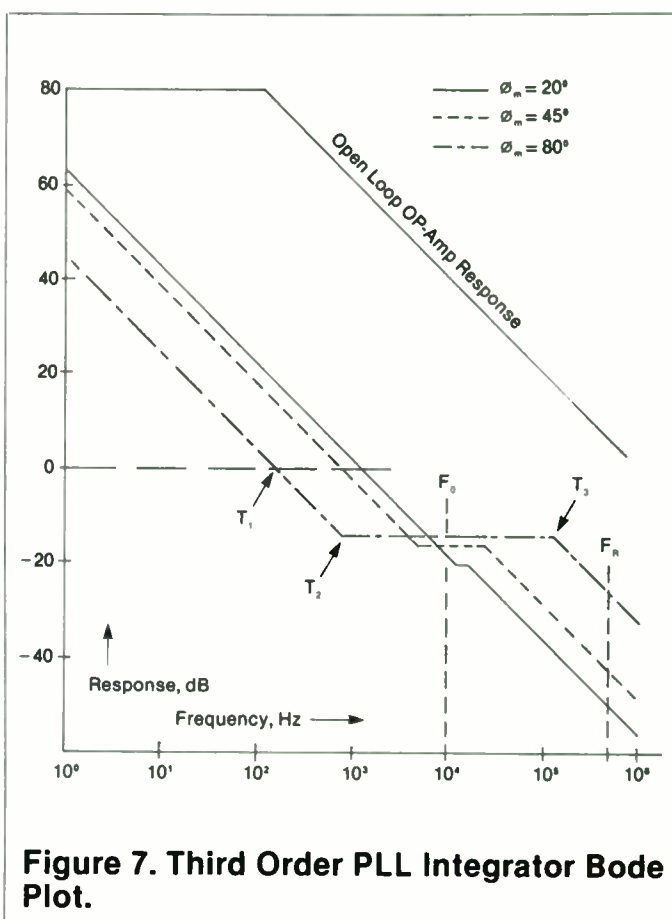




tracking<sup>2, 3</sup>. A typical circuit is shown in Figure 6. To provide good predictable performance, the integrator response curve should fall below the op-amp open loop response, as shown in Figure 7, down to below one Hz and above T3. The three time constants, T1, T2, and T3 define the integrator.

$$\begin{aligned} T_1 &= C_1 \times R_1 & (1) \\ T_2 &= C_1 \times R_2 & (2) \\ T_3 &= C_2 \times R_2 & (3) \end{aligned}$$

They can be easily calculated by using the optimizing technique previously described<sup>7</sup>. This technique is much easier



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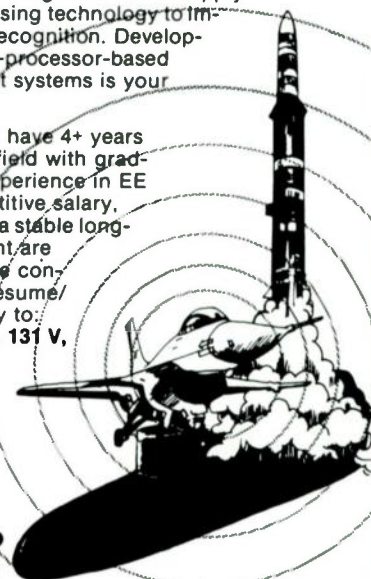
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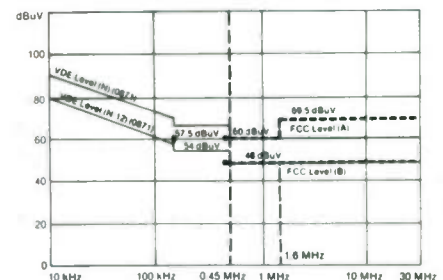
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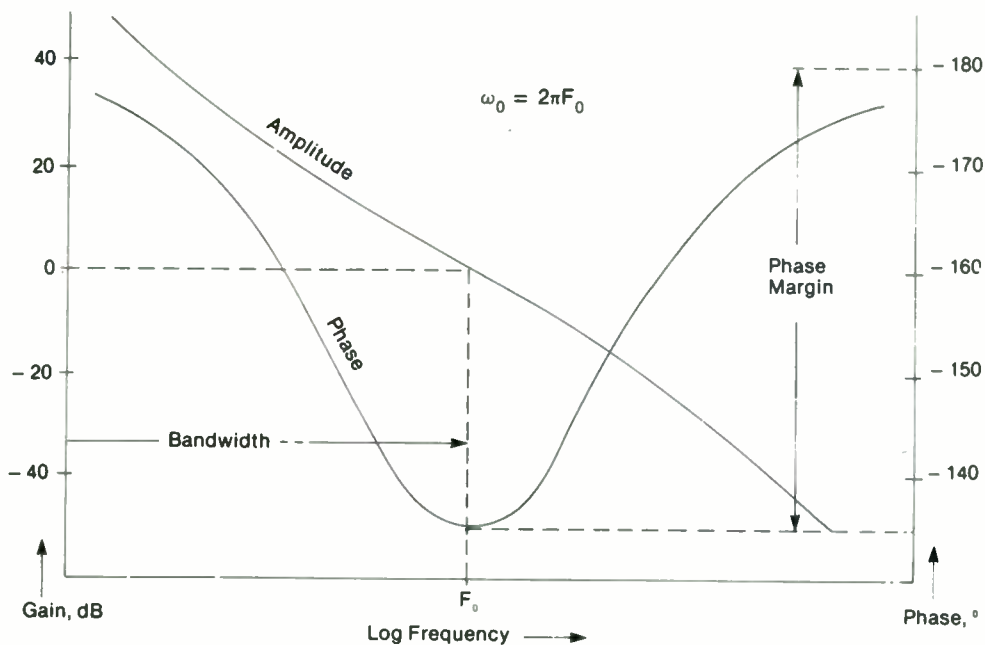
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**Figure 8. Open Loop Response,  $G(s)H(s)$ .**

to use than the second order approximations<sup>2</sup> usually used and gives more predictable results.

$$T3 = \frac{\sec \phi - \tan \phi}{\omega_0} \quad (4)$$

$$T2 = \frac{2}{\omega_0} \tan \phi \quad (5)$$

$$T1 = \frac{K_p K_v}{N \omega_0^2} \sqrt{\frac{[\omega_0(T_2 + T_3)]^2 + 1}{(\omega_0 T_3)^2 + 1}} \quad (6)$$

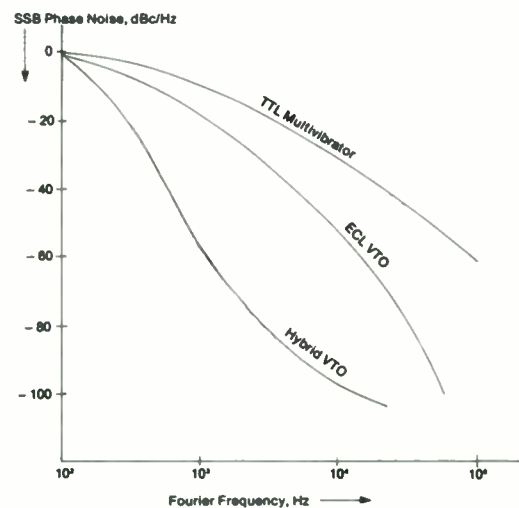
where  $\phi$  = phase margin  
and  $\omega_0$  = loop "bandwidth"  
as defined in Figure 8.

The HP-41 program of Table I simplifies the calculations. In using the three time constants, note that sometimes  $T2$  is defined in the literature as  $R2(C1 + C2)$  or  $T2 + T3$ , as used in this article. Note also that phase margins over  $90^\circ$  cannot be obtained and that  $T2$  and  $T3$  are independent of  $K_v$ ,  $K_p$  and  $N$ . Either a single ended or differential integrator is used depending on the type of phase comparator (Figure 5).

Different bandwidths (such as 3dB, noise, natural frequency) have been used<sup>3</sup> in second order PLL analysis. To facilitate design calculations loop "bandwidth" was defined as 0 to  $\omega_0$  radians, as shown in Figure 8. The cut-off frequency,  $\omega_0$ , is the point at which the open loop gain drops to unity (0 DB). Thus, the phase margin (stability criterion) is the conventional stability margin as used in operational amplifier design<sup>4</sup>.

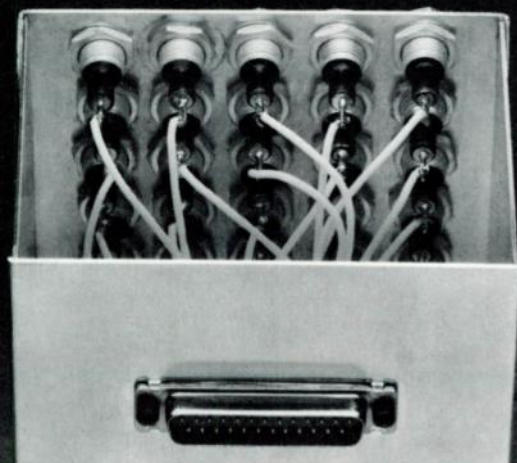
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Several choices exist, depending on frequency, range and noise characteristics. The easiest to use is the monolithic RC-tuned VTO. This type has very large range and its control characteristic is very linear ( $K_v$  is constant). However, its noise characteristics are poor, as seen in Figure 9. The



**Figure 9. VTO Phase Noise Comparison.**

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**Table I**

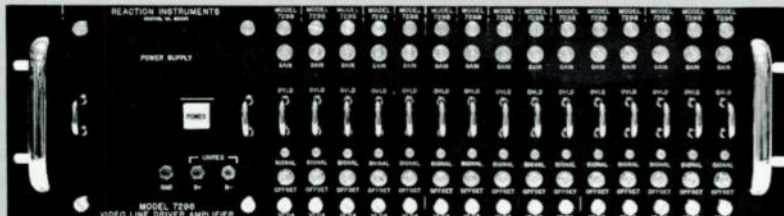
**REGISTERS**

01+LBL "PLL TC"	20 AVIEW	39 *	58 RTN
02 FS? 21	21 RCL 19	40 X↑2	59+LBL 01
03 XEQ 01	22 TAN	41 1	60 "KP = "
04 SF 21	23 RCL 20	42 +	61 ARCL 15
05 RCL 19	24 /	43 /	62 AVIEW
06 COS	25 ST+ X	44 SQRT	63 "KV = "
07 1/X	26 STO 13	45 RCL 15	64 ARCL 16
08 RCL 19	27 "T2 ="	46 *	65 AVIEW
09 TAN	28 ARCL X	47 RCL 16	66 "N = "
10 -	29 AVIEW	48 *	67 ARCL 17
11 RCL 18	30 RCL 14	49 RCL 17	68 AVIEW
12 ST+ X	31 +	50 /	69 "FO = "
13 PI	32 RCL 20	51 RCL 20	70 ARCL 18
14 *	33 *	52 X↑2	71 AVIEW
15 STO 20	34 X↑2	53 /	72 "Δ = "
16 /	35 1	54 STO 12	73 ARCL 19
17 STO 14	36 +	55 "T1 ="	74 AVIEW
18 "T3 ="	37 RCL 14	56 ARCL X	75 ADV
19 ARCL X	38 RCL 20	57 AVIEW	76 END

R01	_____
R02	_____
R03	_____
R04	_____
R05	_____
R06	_____
R07	_____
R08	_____
R09	_____
R10	_____
R11	_____
R12	[T1]_____
R13	[T2]_____
R14	[T3]_____
R15	K <sub>p</sub> _____
R16	K <sub>v</sub> _____
R17	N_____
R18	F <sub>o</sub> _____
R19	φ <sub>m</sub> _____
R20	[ω <sub>0</sub> ]_____
R21	_____
R22	_____
R23	_____
R24	_____

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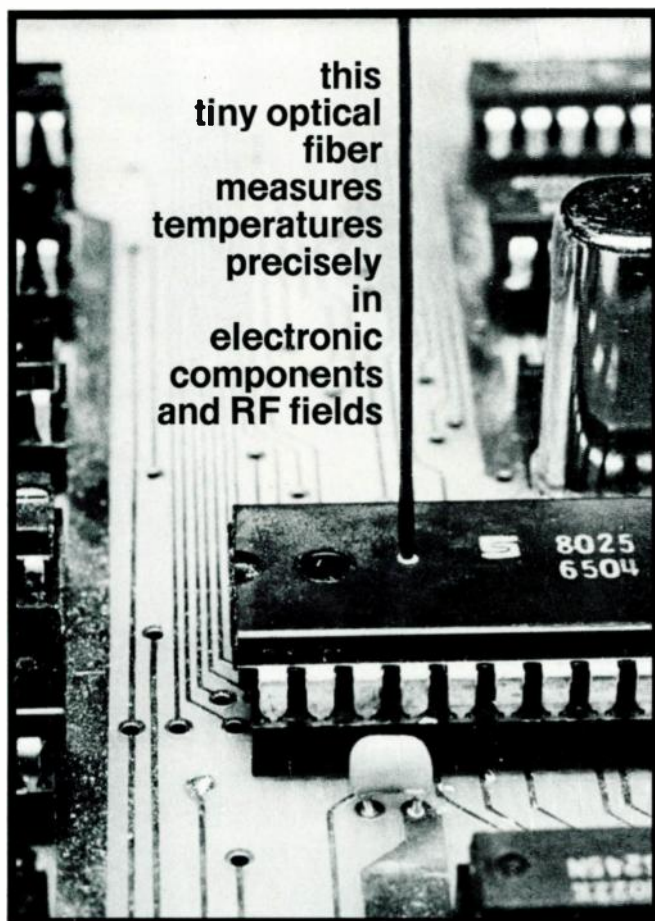
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RC VTO	very good	very high	very large	low
LC ECL	medium	medium	medium	medium
LC Hybrid	medium	low	medium	high
VCXO	poor	very low	very low	low

**Figure 10. VTO Comparison.**

LC-tuned ECL VTO has smaller range, lower noise and poorer linearity. The LC-tuned hybrid or discrete VTO has the lowest noise, but again, lower linearity and smaller tuning range.

In some special low noise, low range applications, a crystal controlled VTO may be used<sup>10</sup>. A comparison of these types is summarized in Figure 10.

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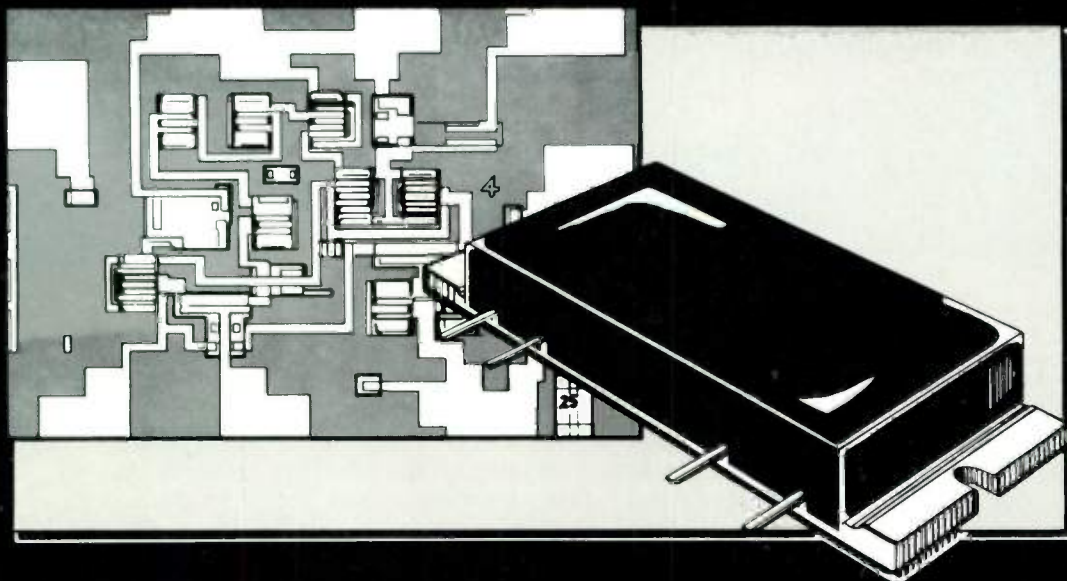
With the exception of the 1:1 tracking loop, some form of frequency division in the feedback path of the loop is needed to provide the required closed loop frequency multiplication. Usually a fixed or programmable digital divider is used. If a programmable divider with a large divide-by-N is used, it is sometimes necessary to use a pulse stretcher to operate the usually low-speed phase comparator.

*Part II of this article to appear in the May/June issue will continue the discussion with design parameters and possible problems. Ed.*

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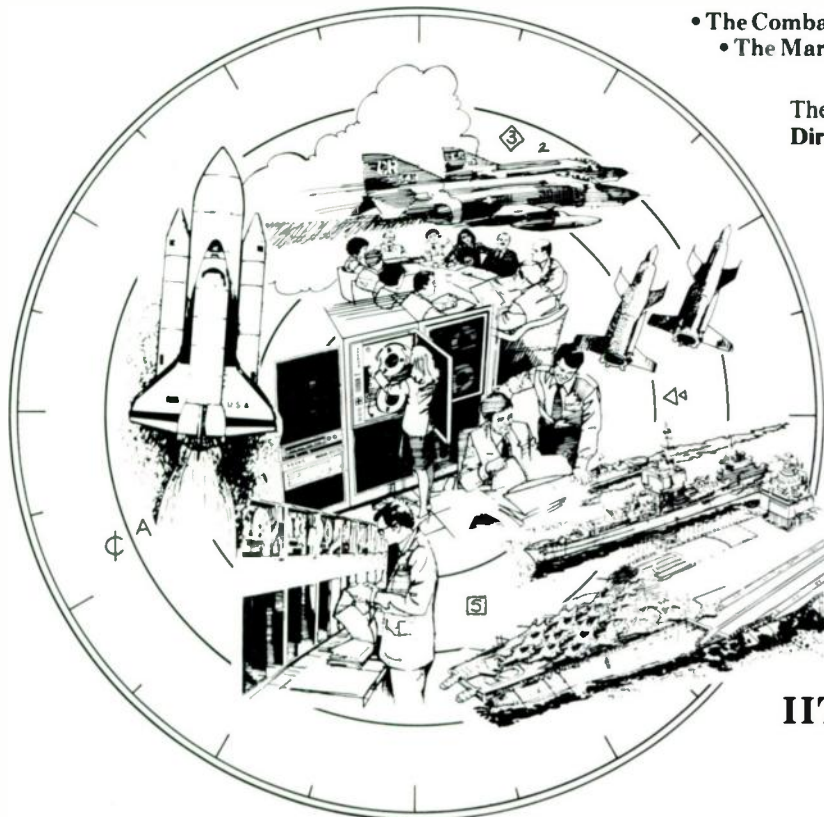
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# RF Amplifier Design Program

A calculator software package able to evaluate RF amplifier performance and determine required source and load terminations for any condition of stability.

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School of Electrical Engineering  
Georgia Institute of Technology  
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**T**he HP-41 calculator is a very flexible and powerful instrument. Its ability to handle conveniently alphanumeric data makes the construction of highly interactive, easy-to-use programs possible. This article presents a package of RF design software using the aforementioned capabilities of the HP-41. This program will design for all conditions of transistor stability and can utilize either  $y$ - or  $S$ -parameters.

## General Description of the Software

This software gives the designer the ability to determine stability factors, amplifier gains, input and output admittances, and terminations for a specified stability factor. A conversion routine is included so that the design may be accomplished with  $S$ -parameters if desired.

The ability of the HP-41 to easily work with alphanumeric data is employed to make all programs interactive. The "customized keyboard" feature is not used, but instead the programs are operated from a "menu" as often done with computers. As a consequence of this technique almost no instructions are needed to use the software. This eliminates many data input errors and is especially time saving if the design programs are used only occasionally.

The software is organized around a main program and eight subroutines. The main program contains the menu and calls appropriate subroutines according to the task selected. Flags keep track of quantities that have been entered or generated, e.g.,  $y$ -parameters or load terminations.

Thus, when a given function is selected, the main program calls only for quantities that have not been entered. If all data are available, the desired calculation is completed.

Upon initial execution of the main program, the calculator prints the menu of possible commands. These are:

- 1 - LSF
- 2 - OPT TERM
- 3 - GAINS
- 4 - YIN/OUT
- 5 - CHNG YS/L
- 6 - CHNG YPAR
- 7 - K-STERN
- 8 - YS/L FOR K
- 9 - S/Y CONV

Commands are executed by entering the appropriate number and pressing R/S. The calculator will prompt the user for any necessary data. Upon execution of any command, the program returns to the command mode displaying the message, CMND. At this time it is ready to execute the function. All entered and calculated data are available for the next operation. The operation of the nine commands will now be summarized.

1. LSF. The first command calculates the Linvill stability factor of the device. It also checks to see if  $\text{Re}(y_o) < 0$  and  $\text{Re}(y_i) < 0$ . If either real part is negative, the device is potentially unstable and the Linvill factor is not a valid indicator. A message indicating this fact is displayed.

2. OPT TERM. This command calculates the optimum terminations ( $Y_s$  and  $Y_L$ ) and the optimum gain for an in-

herently stable transistor. The routine first checks the Linvill stability factor and prints out an appropriate message if the optimum terminations do not exist.

3. GAINS. The third command calculates the available, operating and transducer power gains. If a particular gain does not exist, a message indicating that fact is given.

4. YIN/OUT. This command causes the input and output admittances to be calculated. It requires the y-parameters and the source and load admittances. The user is prompted to enter these data if they are not already present.

5. CHNG YS/L. Command number five allows the user to change the values of the source and load admittances. Some routines generate  $Y_s$  and  $Y_L$ , and some require  $Y_s$  and  $Y_L$ . This routine enables the current values of  $Y_s$  and  $Y_L$  to be changed without disturbing other data in the calculator.

6. CHNG YPAR. This routine is similar to number five, but it is used to change the current values of the y-parameters.

7.K-STERN. This command calculates Stern's stability factor for the circuit. It needs the device parameters and the source and load terminations.

8. YS/L FOR K. Command number eight calculates the required source and load terminations for a specified value of Stern's stability factor. An iterative procedure is used which stops when the change between successive iterations in the imaginary parts of both  $Y_s$  and  $Y_L$  is less than or equal to 0.01 percent. Only the device parameters are needed for this calculation.

9. S/Y CONV. The final routine converts S-parameters to y-parameters. Normalized 50-ohm S-parameters are assumed, which are entered in polar form. The program calculates the y-parameters and loads them into the appropriate storage locations. Once the S-parameters have been entered, any of the various commands may be executed.

## Preliminary Operations

Before the program can be utilized, it must be loaded into the calculator. The complete program occupies 1821 bytes of code and requires a minimum SIZE of 043. With a SIZE of 043, there will be 17 registers free when all routines are loaded. If S-parameters will not be used the SPAR routine and associated storage will not be needed. In this case only 1512 bytes are used and the minimum SIZE of 030 will leave 74 registers unused. None of the routines use registers 00 and 01 so these are free for temporary storage while using the software. Ten (10) complete magnetic cards are required to store the entire package.

## Using the Software

Once loaded, the only other factor to be considered is the presence of the HP82143A printer. The software works best with the printer, but it is completely usable without the device. If the printer is not present, it is recommended that flag no. 21 be set manually before using the programs. With flag no. 21 set, the calculator will stop whenever a print or prompt command is executed. Pressing R/S will continue the program. If flag no. 21 is not set, the calculator rapidly goes through the explanatory statements, but stops on all prompts. If the printer is used, it should be set to the NORMAL mode.

## Examples of Program Use

The utility and flexibility of the software will be illustrated by two examples. The first example uses y-parameters and an inherently stable transistor while the second considers a transistor described by S-parameters that is not inherently stable. All features of the programs are illustrated by these examples.

r.f. design

## Example I.

This example illustrates several aspects of the software by focusing on an inherently stable transistor specified by y-parameters. The method of design will be discussed in detail with a printout of all entered and generated data listed on the right of the discussion.

Given a 2N5109 transistor operated at 200 MHz. The y-parameters in mmho for the transistor are:

$$\begin{aligned} y_{ie} &= 22 + j9 & y_{re} &= 0 - j2.2 \\ y_{fe} &= 40 - j185 & y_{oe} &= 1 + j8 \end{aligned}$$

Start the design procedure by executing program AMP. The program displays the command table and requests a command number. Enter 1 to evaluate the Linvill stability factor. The y-parameters are requested. Enter them as requested, pushing R/S each time. The Linvill stability factor is computed and printed. Since it is less than one, the transistor is inherently stable.

After completing the stability factor calculation, the calculator requests the next command. Enter 2 to calculate the optimum load and source admittances. The optimum termination routine always calculates the Linvill stability factor in addition to the optimum terminations and gain. Knowing this, the first command could have been omitted.

Any desired command can now be executed. For example, what are the input and output admittances? Run command No. 4. The result confirms that a conjugate match exists at both ports of an optimally terminated, inherently stable device.

Suppose that the transistor will not be optimally terminated, but that

$$Y_s = 20 - j50 \text{ mmho}$$

and

$$Y_L = 1.5 + j5 \text{ mmho}$$

What will be the various gains under these conditions?

It is first necessary to change the current values of  $Y_s$  and  $Y_L$ . (Remember, the optimum values of  $Y_s$  and  $Y_L$  are still stored in the calculator.) To change  $Y_s$  and  $Y_L$ , execute command no. 5. After entering the new values of  $Y_s$  and  $Y_L$ , execute routine no. 3 which will calculate all the power gains. Since all the gains are different, neither of the ports is conjugately matched.

The calculator is ready to receive the next command, but it isn't going to get one. This completes Example I.

## Example II.

The second example will be done with S-parameters to illustrate how they are handled by the program. In addition, the device used is potentially unstable, so other aspects of the software are demonstrated.

Given a 2N6680 GaAs FET operated at 4 GHz. The 50-ohm S-parameters are:

$$\begin{aligned} S_{11} &= 0.792 \angle -107.4 & S_{12} &= 0.070 \angle 26 \\ S_{21} &= 1.995 \angle 81.1 & S_{22} &= 0.659 \angle -60.8 \end{aligned}$$

A. Assume that it is desired to design an amplifier having at least 12 dB transducer power gain with adequate stability.

Start the design procedure by executing the design program AMP. The program displays the command table and requests a command.

Since this design will be done with S-parameters, command no. 9 should be entered. The program requests the S-parameters which will be converted to y-parameters and a flag set indicating that the y-parameters are present.

Guess that the transistor is inherently stable. Enter

	XEQ "AMP"	GR=?		GO=?			
		0	RUN		1-03	RUN	BS=?
ENTR CMND		BR=?		BO=?			-50-03 RUN
1-LSF		-2.2-03	RUN		8-03	RUN	GL=?
2-OPT TERM		CMND?					1.5-03 RUN
3-GAINS		2.	RUN	L. S. F. = 923.3E-3			BL=?
4-YIN/OUT							5-03 RUN
5-CHNG YS/L		OPTIMUM VALUES		CMND?			
6-CHNG YPAR				4.	RUN	CMND?	
7-K-STERN		L. S. F. = 923.3E-3		GIN= 86.61E-3			3. RUN
8-YS/L FOR K				BIN= 53.00E-3			
9-S/Y CONV		GS= 86.61E-3				OPERATING GAIN	
		BS= -53.00E-3		GOUT= 3.937E-3		GP= 9.509E0 dB	
CMND?				BOUT= 10.00E-3			
	1.	GL= 3.937E-3				AVAILABLE GAIN	
	RUN	BL= -10.00E-3				GA= 16.27E0 dB	
ENTER Y-PARAMS				CMND?			
		GMAX= 17.59E0 dB		5.	RUN	TRANSDUCER GAIN	
GI=?		GF=?				GT= 4.952E0 dB	
	22-03		40-03 RUN	ENTER YS AND YL			
BI=?		BF=?		GS=?		CMND?	
	9-03		-135-03 RUN		20-03	RUN	

### Example 1.

	XEQ "AMP"	CMND?		AVAILABLE GAIN		ENTER YS AND YL
		2.	RUN	GA= 13.74E0 dB		GS=?
ENTR CMND						10.22-03 RUN
1-LSF				TRANSDUCER GAIN		BS=?
2-OPT TERM		OPTIMUM VALUES		GT= 12.77E0 dB		-20.23-03 RUN
3-GAINS		L. S. F. = 1.268E0				GL=?
4-YIN/OUT				CMND?		6.749-03 RUN
5-CHNG YS/L		DO NOT EXIST		5.	RUN	BL=?
6-CHNG YPAR						-14.29-03 RUN
7-K-STERN				ENTER YS AND YL		
8-YS/L FOR K				GS=?		CMND?
9-S/Y CONV					10.22-03	7. RUN
		CMND?		BS=?		
CMND?		8.	RUN	-20.23-03	RUN	K-STERN= 6.457E0
	9.	K=?		GL=?		
MS11=?		10	RUN		1-03	CMND?
	.792			BL=?		3. RUN
AS11=?		GS= 12.60E-3			1-03	
	-107.4	BS= -29.75E-3				
MS12=?				CMND?		OPERATING GAIN
	.070	GL= 10.04E-3		4.	RUN	GP= 14.40E0 dB
AS12=?		BL= -14.74E-3				AVAILABLE GAIN
	26			GIN= 10.58E-3		GA= 11.85E0 dB
MS21=?				BIN= 23.65E-3		
	1.995	CMND?		GOUT= 6.749E-3		TRANSDUCER GAIN
AS21=?		3.	RUN	BOUT= 14.29E-3		GT= 11.85E0 dB
	81.1					
MS22=?		OPERATING GAIN		CMND?		CMND?
	.659	GP= 13.74E0 dB				
AS22=?				5.	RUN	
	-60.8					

### Example 2.

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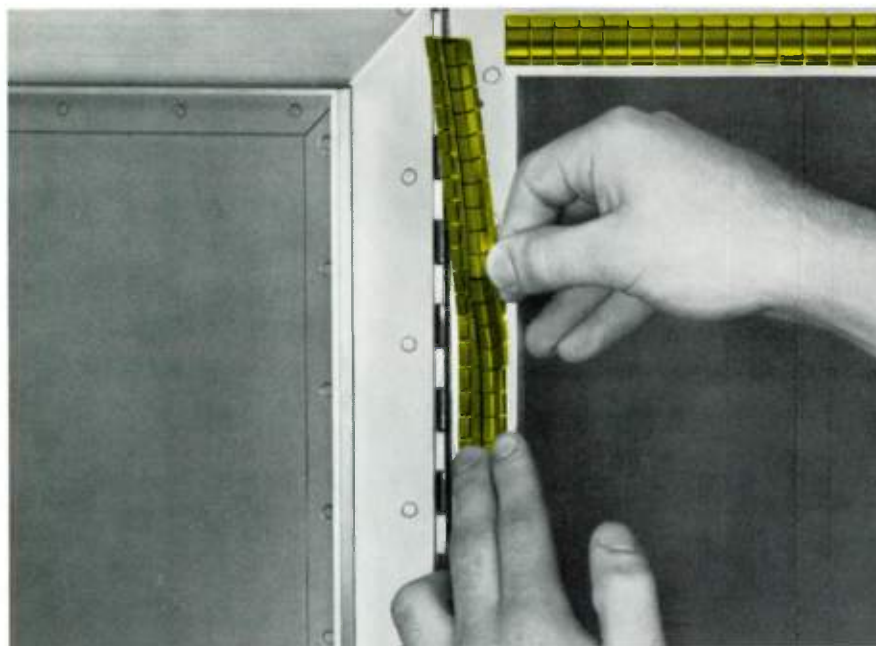
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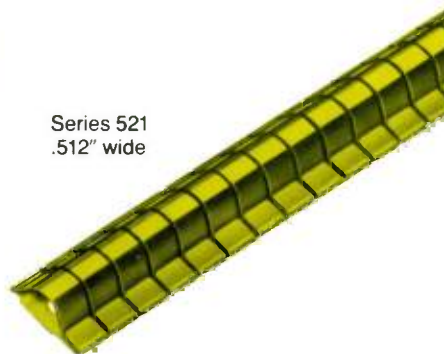
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command no. 2 to calculate the optimum terminations. The calculator informs the user that the device is not inherently stable and consequently the optimum terminations do not exist.

Since the device is not inherently stable, Stern's procedure can be used to stabilize the circuit to any desired degree. The choice of stability factor affects the gain. Typically, the stability factor should be greater than about five. Select an initial stability factor of 10 and execute routine no. 8 to generate the  $Y_s$  and  $Y_L$  given that value of  $k$ . The iterative procedure takes about 56 seconds.

How much gain will be realized? Execute command no. 3 to calculate all of the gains. The amplifier will deliver a transducer power gain of 12.77 dB which meets the specifications. The stability factor is quite large, so some of the stability could be traded for additional gain.

B. Rather than do another example of Stern's procedure for a smaller stability factor, assume that it is required that the amplifier have minimum noise figure.

The noise figure of an amplifier is a function of the source admittance. The 2N6680 transistor has a minimum noise figure of 1.60 dB with a source reflection coefficient of 0.618/98. This corresponds to a source admittance of  $(10.22 - j20.23)$  mmho. Specifying the source termination sets the available power gain. In order to realize an overall or transducer power gain equal to the available power gain, the output of the transistor must be conjugately matched.

Begin in the design by calculating the output admittance of the amplifier when the input is terminated with the source admittance for best noise figure. It is necessary to change the current value of the source admittance by executing command no. 5.

Enter the value of  $Y_s$  for best noise figure. Since the program requests  $Y_L$  also, enter some dummy value.

Next, execute command no. 4 to find out what  $Y_{out}$  will

be when the input is adjusted for minimum noise figure. The calculated value of the input admittance, which is a function of the dummy value of load impedance, is ignored.

Use the change  $Y_s/Y_L$  routine again, but this time make  $Y_L = Y_{out}$ . Now, both terminations are entered in the calculator and other quantities may be determined.

First, check the stability of the circuit by executing command no. 7. The  $k$ -factor of 6.457 is adequate for good stability.

As the last step, execute command no. 3 which determines the transducer gain of the optimized amplifier. A transducer power gain of 11.85 dB is obtained with a noise figure of 1.60 dB. This completes example II.

## Summary

Calculator software able to evaluate amplifier performance and to determine required source and load terminations for any condition of stability has been developed. A complete listing of the program, together with the applicable equations<sup>1,2</sup> is presented in Appendix I. Appendix II contains a listing of data storage assignments. Given this information, individual subroutines can be included in user-developed programs. □

## References

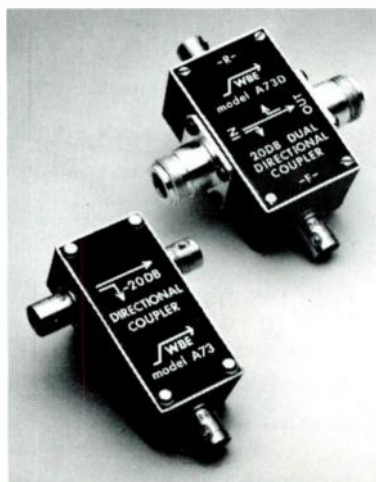
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Model	Freq Range MHz	Coupler Type	In Line Power	Minimum Directivity (dB)		In Line Loss (dB)	Response Flatness of -20 dB part (dB)	VSWR
				1-300 MHz	3-300 MHz			
A73-20	1-500	single	5W cw (10W cw 5-300 MHz)	20	30	.4 max .2 typical	±.1 5-300 MHz ±.25 1-500 MHz	1.1:1 5-500 1.5:1 1-500
A73-20GA				30	40			
A73-20GB				40	45			
A73-20P	1-100	single	50W cw (75 ohm limited to 10W cw)	35 dB min		.15	±.1	1.1:1 max
A73D-20P		dual		40 dB min typical		.3		
A73-20PX		single		45 dB min		.15		
A73D-20PX		dual		45 dB min		.3		
A73-20PA	10-200	single		35 dB min		.15		1.04:1 typical
A73D-20PA		dual		40 dB min typical		.3		
A73-20PAX		single		45 dB min		.15		
A73D-20PAX		dual				.3		

## WIDE BAND ENGINEERING COMPANY, INC.

P O BOX 21852, PHOENIX, ARIZONA 85036, U S A • TELEPHONE (602) 254-1570

INFO/CARD 27

INFO/CARD 26

REF LEVEL

INPUT ATTEN

LOG SCALE

10 dBm

10 dB ATTEN

10 dB

When you're shooting for 99.95% availability on your satellite circuits, you don't gamble. RCA Americom use PTS 160 Synthesizers in their Earth Stations; from Hawaii to Greenland they have shown excellent reliability and performance for years.

If you need VHF-UHF Synthesizers for precision frequency control, specify PTS.

**PTS**

FREQUENCY SYNTHESIZER MODELS  
40 MHz, 160 MHz, 200 MHz, 500 MHz


**PTS**

PROGRAMMED TEST SOURCES, INC.

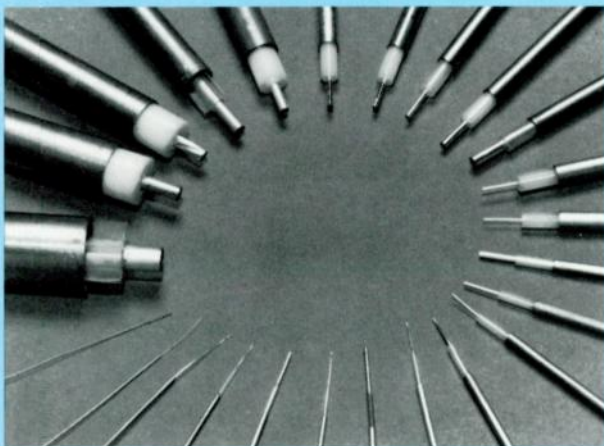
Littleton, MA, 617-486-3008  
INFO/CARD 28

# Cost Cutting Hints For Semi-Rigid Coaxial Cables

**HINT #1:** Keep it simple. Using standard, rather than special semi-rigid coaxial cables saves you time and money while providing the transmission characteristics you need. **Uniform offers 66 standard semi-rigid cables** that meet the requirements of nearly all normal applications without going into special designs. Twelve meet established MIL-C-17E specs.

**HINT #2:** Consider using short random lengths under five feet wherever practical. Frequently long lengths are not required and first quality shorter lengths can **save up to 30%** when they are available. All it takes to find out is a phone call. Lengths under five feet can often be sent to you within three days of your call.

**HINT #3:** Consider ordering complete high-reliability cable assemblies pretested to your specifications if required. **This saves time and money** in procurement procedures, inspection and testing, and manufacturing and delivery scheduling.



Whatever your requirements in semi-rigid coax from 500 MHz to 40 GHz, and diameters from 0.008" (0.02MM) to 0.50" (12.7MM), Uniform Tubes remains the reliable source. Can we supply specials? Of course...and in all sizes and characteristics to your specific requirements.

Send for informative Bulletin 724 on Micro-Coax® Semi-Rigid Coaxial Cables, Cable Assemblies, and Coaxial and Waveguide Delay Lines.



**MicroDelay** / **UNIFORM TUBES, INC.**  
Division ...a UTI company

COLLEGEVILLE, PA 19426, U.S.A. • 215/539-0700  
TWX 510-660-6107 • Telex 84-6428

Represented in Europe by Huber + Suhner AG, Hensau, Switzerland • Telex 77-426

INFO/CARD 29

## Appendix I. Program Listing.

### 1. Main Program

This program contains the command menu and calls subroutines required to execute a specific command.

01•LBL "AMP"	35 ADV	69 "ROUT="
02 CF 04	36 XEQ "LINV"	70 ARCL 24
03 CF 03	37 GTO 00	71 AVIEW
04 ADV	38•LBL 02	72 ADV
05 "ENTR CMND"	39 FC? 03	73 GTO 00
06 AVIEW	40 XEQ "PARA"	74•LBL 05
07 "1-LSF"	41 XEQ "OPT"	75 XEQ "YS/L"
08 AVIEW	42 GTO 00	76 GTO 00
09 "2-OPT TERM"	43•LBL 03	77•LBL 00
10 AVIEW	44 FC? 03	78 XEQ "PARA"
11 "3-GAINS"	45 XEQ "PARA"	79 GTO 00
12 AVIEW	46 FC? 04	80•LBL 07
13 "4-YIN/OUT"	47 XEQ "YS L"	81 FC? 03
14 AVIEW	48 XEQ "YIN"	82 XEQ "PARA"
15 "5-CHNG YS/L"	49 XEQ "SP"	83 FC? 04
16 AVIEW	50 XEQ "SP"	84 XEQ "YS/L"
17 "6-CHNG YPAR"	51 XEQ "GT"	85 XEQ "STN"
18 AVIEW	52 GTO 00	86 ADV
19 "7-K-STERM"	53•LBL 04	87 "K-STERM="
20 AVIEW	54 FC? 03	88 ARCL 17
21 "8-YS/L FOR K"	55 XEQ "PARA"	89 AVIEW
22 AVIEW	56 FC? 04	90 GTO 00
23 "9-S/Y CONV"	57 XEQ "YS/L"	91•LBL 08
24 AVIEW	58 XEQ "YIN"	92 FC? 03
25•LBL 00	59 "GIN="	93 XEQ "PARA"
26 ADV	60 ARCL 23	94 XEQ "ITR"
27 FIX 0	61 AVIEW	95 ADV
28 "CMND?"	62 "BIN="	96 XEQ "S/LPR"
29 PROMPT	63 ARCL 22	97 GTO 00
30 ENG 3	64 AVIEW	98•LBL 09
31 GTO IND X	65 ADV	99 XEQ "SPAR"
32•LBL 01	66 "GOUT="	100 XEQ "PAR1"
33 FC? 03	67 ARCL 25	101 GTO 00
34 XEQ "PARA"	68 AVIEW	102 END

### 2. Linvill Stability Factor

Applicable Equations:

$$C = \frac{|y_r y_l|}{2g_o - \operatorname{Re}(y_r y_l)}$$

Flags:

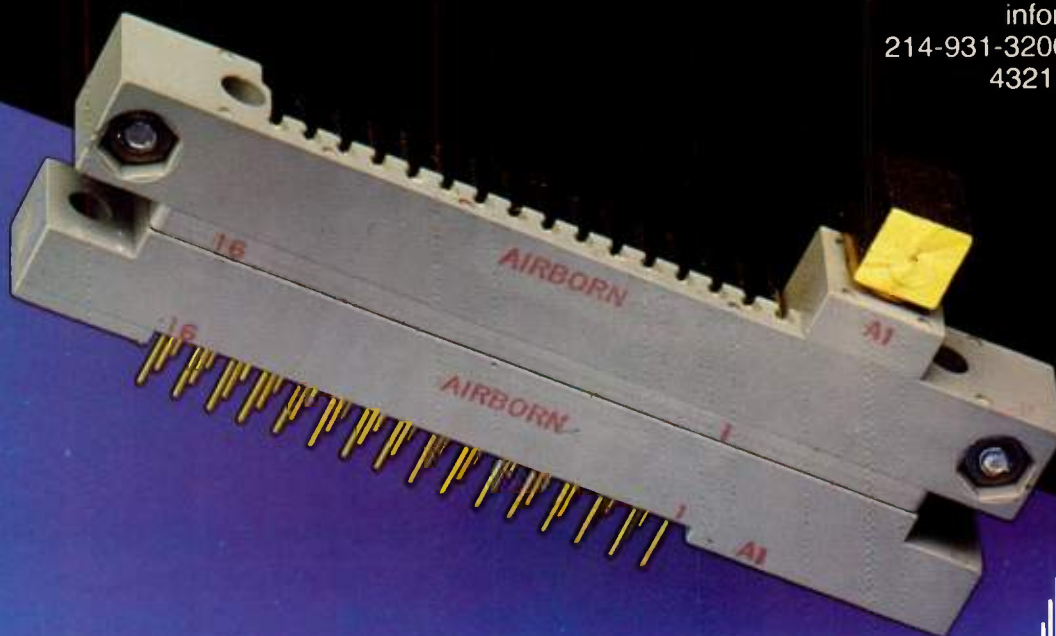
00 Set if  $g_l < 0$  or  $g_o < 0$  or  $C < 0$  or  $C \geq 1$ .

01•LBL "LINV"	11 /	21 AVIEW
02 CF 00	12 STO 16	22 ADV
03 RCL 03	13 X<0?	23 RTN
04 X<0?	14 SF 00	24•LBL 01
05 GTO 01	15 1	25 SF 00
06 RCL 05	16 X<>Y	26 "UNSTABLE"
07 X<0?	17 X>Y?	27 AVIEW
08 GTO 01	18 SF 00	28 "G/O/O/O"
09 RCL 12	19 "L. S. F. =	29 AVIEW
10 RCL 14	20 ARCL X	30 END

# 47 PLUS 1 COAX



AirBorn offers you a combination connector for PC Board application which has one coaxial contact and 47 DC contacts\*. 50-ohm characteristic impedance. For any application, get perfect integration of DC and Coax signals with complete circuit integrity. For more information and specifications, call 214-931-3200; or write AirBorn Connectors, 4321 AirBorn Drive, Addison, Texas 75001.



\*DC contacts are size 22 with staggered terminal spacings of .100" x .125" (plugs) and .100" x .100" (receptacles).



INFO/CARD 30

## Appendix I. Program Listing . . . Con't.

### 3. Optimum Terminations

Applicable Equations:

$$P = \operatorname{Re}(y_i y_i)$$

$$Q = \operatorname{Im}(y_i y_i)$$

$$M = |y_i y_i|$$

$$G_{S_{\text{opt}}} = \frac{1}{2g_o} \sqrt{(2g_i g_o - P)^2 - M^2}$$

$$B_{S_{\text{opt}}} = -b_i + \frac{Q}{2g_o}$$

$$G_{L_{\text{opt}}} = \frac{1}{2g_i} \sqrt{(2g_i g_o - P)^2 - M^2}$$

$$B_{L_{\text{opt}}} = -b_o + \frac{Q}{2g_i}$$

Flags:

04 set upon successful completion to indicate  $Y_S$  and  $Y_L$  generated.

01 LBL "OPT"	15 STO 15	25 /
02 ADV	16 RCL 05	30 2
03 "OPTIMUM VALUES"	17 /	31 /
04 AVIEW	18 2	32 RCL 04
05 ADV	19 /	33 -
06 XEQ "LINV"	20 STO 19	34 STO 20
07 FS? 00	21 RCL 15	35 RCL 16
08 GTQ 01	22 RCL 03	36 RCL 05
09 RCL 14	23 /	37 /
10 X12	24 2	38 2
11 RCL 12	25 /	39 /
12 X12	26 STO 21	40 RCL 02
13 -	27 RCL 10	41 -
14 SORT	28 RCL 03	42 STO 16

43 XEQ "S/LPK"	58 AVIEW	73 "dB"
44 GTQ 03	59 ADV	74 ASTO Y
45 LBL "S/LPK"	60 RTN	75 "GMAX="
46 "GS="	61 LBL 03	76 ARCL X
47 ARCL 19	62 RCL 08	77 ARCL Y
48 AVIEW	63 RCL 09	78 AVIEW
49 "BS="	64 R-P	79 SF 04
50 ARCL 18	65 X12	80 GTQ 02
51 AVIEW	66 RCL 14	81 LBL 01
52 ADV	67 RCL 15	82 "DO NOT EXIST"
53 "GL="	68 +	83 AVIEW
54 ARCL 21	69 /	84 LBL 02
55 AVIEW	70 LOG	85 ADV
56 "BL="	71 10	86 ADV
57 ARCL 28	72 *	87 .END.

### 4. Power Gains

Applicable Equations:

$$G_A = \frac{|y_i|^2 G_S}{|y_i + Y_S|^2 \operatorname{Re}\{y_o - \frac{y_i y_i}{y_i + Y_S}\}}$$

$$G_P = \frac{|y_i|^2 G_L}{|y_o + Y_L|^2 \operatorname{Re}\{y_i - \frac{y_i y_i}{y_o + Y_L}\}}$$

$$G_T = \frac{4|y_i|^2 G_S G_L}{|(y_i + Y_S)(y_o + Y_L) - y_i y_i|^2}$$

Flags:

None affected.

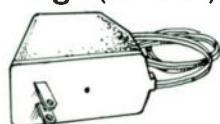
# There are 9 new features on this THRULINE® RF Wattmeter. And one familiar one: the Elements

Forward Power in watts — with 120% overrange and with the decimal point in place.  
The same for Reflected Power

Push for Peak Envelope Power of SSB, AM in Forward or Reflected directions in watts

Push to read CW power in dBm

Portable or AC Line operation with battery charger (included)



# BIRD 4381

0.45-2300 MHz / 0.1-10,000 watts

30303 Aurora Rd. Cleveland (Solon) Ohio 44139  
216 • 248-1200 TLX: 98-5298 Cable: BIRDELEC



Uses the same Plug-In Elements you may already own with your model 43 Wattmeter

Push here to read SWR instantaneously and watch it change

...and % Modulation

...and Return Loss in dB

and while tuning or tweaking, the memory buttons recall min and max values of any chosen quantity—and  $\Delta$  shows if it is rising or falling

**AND NOW: Rackmounted models for 2-way mobile, FM and TV broadcasters, portables with built-in or remote line sections**

**Write for Series 4380  
RF Power ANALYST™**

INFO/CARD 31

01*LBL "GA"	37 RCL 21	73 RCL 21
02 ADV	38 XEQ "Z+"	74 XEQ "Z+"
03 "AVAILABLE GAIN"	39 XEQ "MUL"	75 RCL 27
04 AVIEW	40 RCL 21	76 RCL 26
05 FS? 02	41 *	77 XEQ "Z+"
06 GTO 01	42 RCL 23	78 RCL 10
07 RCL 02	43 /	79 RCL 11
08 RCL 03	44 XEQ "DB"	80 XEQ "Z-"
09 RCL 18	45 "GP= "	81 XEQ "MUL"
10 RCL 19	46 GTO 02	82 RCL 19
11 XEQ "Z+"	47 RTN	83 *
12 XEQ "MUL"	48*LBL "GT"	84 RCL 21
13 RCL 19	49 ADV	85 *
14 *	50 "TRANSDUCER GAIN"	86 4
15 RCL 25	51 AVIEW	87 *
16 /	52 RCL 19	88 XEQ "DB"
17 XEQ "DB"	53 RCL 23	89 "GT= "
18 "GA= "	54 +	90 GTO 02
19*LBL 02	55 XEQ?	91*LBL "DB"
20 ARCL X	56 GTO 01	92 LOG
21 ARCL 28	57 RCL 21	93 10
22 AVIEW	58 RCL 25	94 *
23 RTN	59 +	95 "dB"
24*LBL 01	60 XEQ?	96 ASTO 28
25 "DOES NOT EXIST"	61 GTO 01	97 RTN
26 AVIEW	62 RCL 02	98*LBL "MUL"
27 RTN	63 RCL 03	99 XEQ "ZABS1"
28*LBL "GP"	64 RCL 18	100 X12
29 ADV	65 RCL 19	101 1/X
30 "OPERATING GAIN"	66 XEQ "Z+"	102 RCL 08
31 AVIEW	67 STO 26	103 RCL 09
32 FS? 01	68 XEQ	104 XEQ "ZABS1"
33 GTO 01	69 STO 27	105 X12
34 RCL 04	70 RCL 04	106 *
35 RCL 05	71 RCL 05	107 END
36 RCL 20	72 RCL 20	

## 5. Input and Output Admittance Applicable Equations:

$$Y_{IN} = y_i - \frac{y_r y_l}{y_o + Y_L}$$

$$Y_{OUT} = y_o - \frac{y_r y_l}{y_i + Y_S}$$

### Flags:

01 set if  $\text{Re}(Y_{IN}) < 0$

02 set if  $\text{Re}(Y_{OUT}) < 0$

01*LBL "Y1/0"	18 ISG 26	35 RCL IND 29
02 CF 01	19 RCL IND 26	36 DSE 29
03 CF 02	20 RCL IND 27	37 XEQ
04 1.005	21 ISG 27	38 XEQ "Z+"
05 STO 26	22 RCL IND 27	39 STO IND 28
06 18.023	23 ISG 27	40 DSE 28
07 STO 27	24 XEQ "Z+"	41 XEQ
08 25.019	25 XEQ "ZIN"	42 STO IND 26
09 STO 28	26 RCL 10	43 DSE 28
10 5.900	27 RCL 11	44 GTO 01
11 STO 29	28 XEQ "Z+"	45*LBL 03
12*LBL 01	29 CHS	46 RCL 23
13 ISG 26	30 XEQ	47 XEQ?
14 GTO 02	31 CHS	48 SF 01
15 GTO 03	32 XEQ	49 RCL 25
16*LBL 02	33 RCL IND 29	50 XEQ?
17 RCL IND 26	34 DSE 29	51 SF 02
		52 END

# For a power resistor that stays non-X up to vhf, there's only one choice.

The Carborundum® Type SP. Only Carborundum has a ceramic power resistor that behaves like a pure resistance rather than an inductor and/or capacitor. It operates from low audio frequencies up into the vhf range. Each unit is a solid body of resistive material. No windings, no film. Ideal for frequency-sensitive rf applications like feedback loops.

And it gives you extremely high power density, with great surge-handling capability because it's solid.

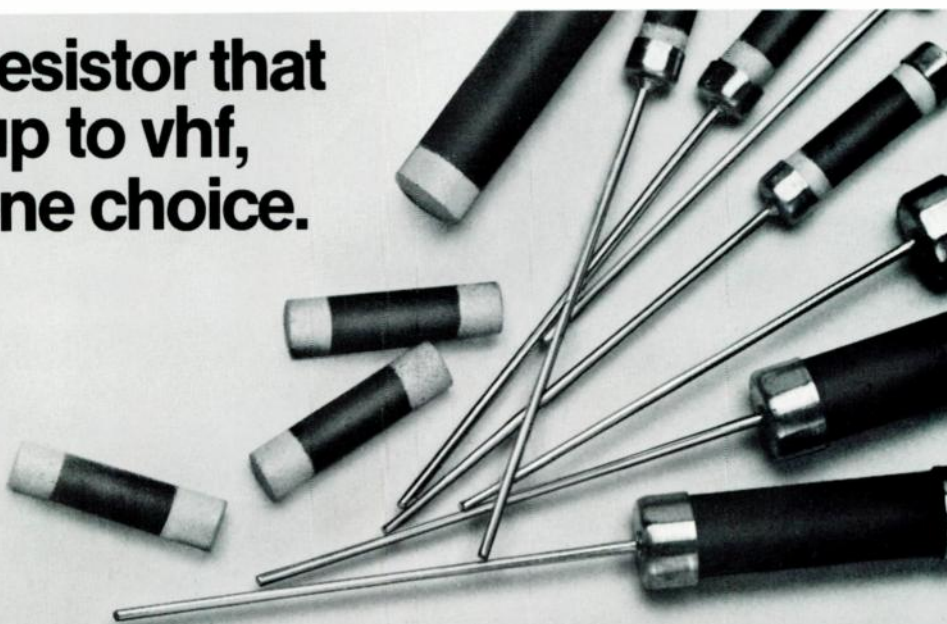
Our Type 234SP, for example, is about the size of a 2-watt carbon comp, but dissipates a full 10 watts in 40°C ambient air. Moreover, it can consistently absorb surges of over 10X rated power for several seconds and come back for more with very little  $\Delta R$ . Forced-air-cooled, water-cooled or

immersed in oil, it will handle even greater power overloads.

Other Carborundum Type SP resistors—including high-power, water-cooled configurations—are rated from 2.5 to 1000 watts. For further details, call or write E. B. (Woody) Hausler at (716) 278-2143.

**Carborundum**  
**Resistant Materials Company**  
Electric Products Division  
P.O. Box 339  
Niagara Falls, New York 14302

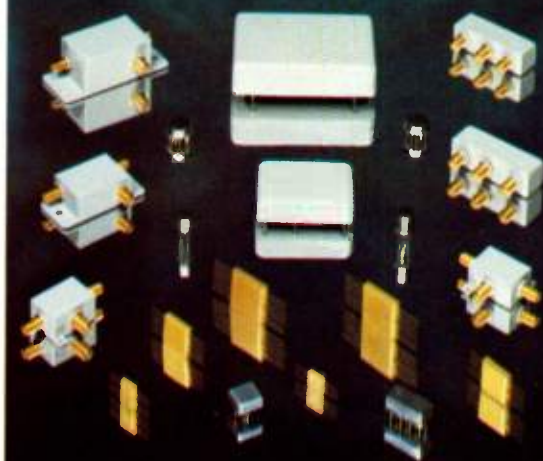
**CARBORUNDUM**  
A Sohio Company  
INFO/CARD 32



# THE SIGNAL PROCESS NOW THE

## IF – BASEBAND COMPONENTS

### MIXERS MODULATORS

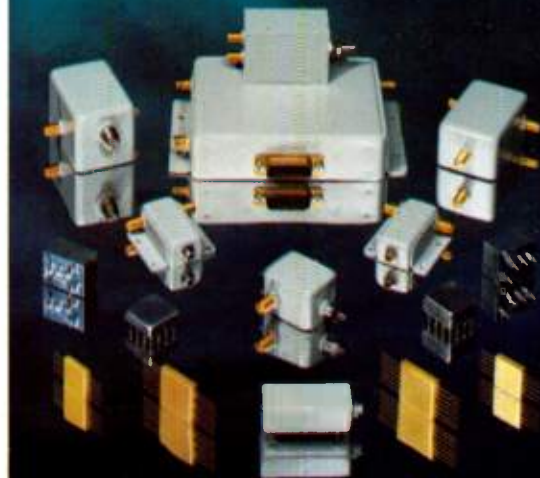


#### FEATURES:

- Frequency Coverage DC - 4.5GHz
- 7 Double Balanced Mixer Families
- LO Drives 0; dBm to + 27 dBm
- Biphas & Quadrphase Modulators
- Frequency Doublers
- 17 Standard Packages

INFO/CARD 33

### VARIABLE PHASE SHIFTERS



#### FEATURES:

- Frequency Coverage 1 - 500 MHz
- Voltage Controlled 0-180°, 360°
- Manually Controlled 0-90°, 180°, 360°
- Digitally Controlled 5, 6, 8 BIT TTL
- BNC, SMA, PC Board Connections
- Panel & Non-Panel Mountable

INFO/CARD 34

 **Merrimac**  
INDUSTRIES, INCORPORATED

P.O. BOX 986 41 FAIRFIELD PLACE WEST CALDWELL, N.J. 07006  
201 575 1300 • TWX 710 734 4314 • TELEX 6853128

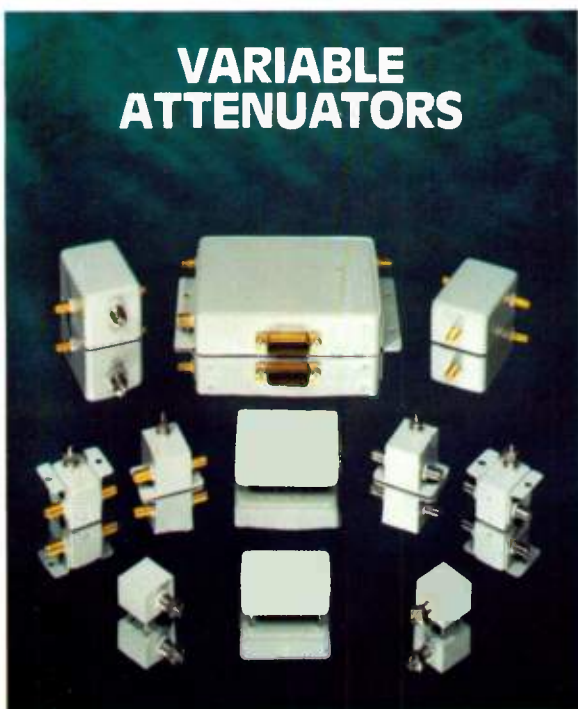
320 PG. IF – BASEBAND  
CATALOG M80-3

INFO/CARD 37

# SING SPECIALIST. RE IS ONE

## DC – 2GHz • 750 CATALOG ITEMS

### VARIABLE ATTENUATORS

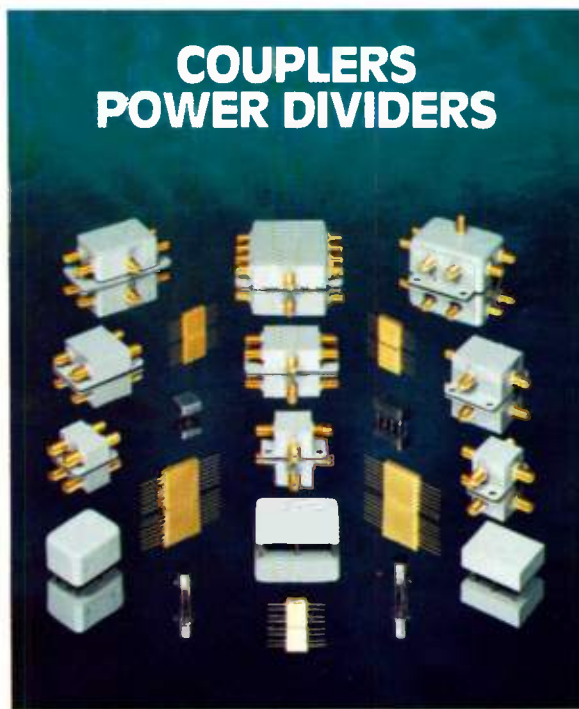


#### FEATURES:

- Frequency Coverage DC - 400 MHz
- Current Controlled 0-20, 30, 40 dB
- Manually Controlled 0-20, 40 dB
- Digitally Controlled 4, 6, 8 BIT TTL
- BNC, SMA, PC Board Connections
- Panel & Non-Panel Mountable

INFO/CARD 35

### COUPLERS POWER DIVIDERS



#### FEATURES:

- Frequency Coverage DC - 2.0 GHz
- 90° 3 dB Quadrature Hybrids
- 0° In-Phase "N"-Way Dividers
- 0° / 180° Hybrid Junctions
- Directional Couplers 6, 10, 20, 30 dB
- 60 Product Families

INFO/CARD 36

130 PG. RF – MICROWAVE  
CATALOG M78-3  
INFO/CARD 38

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201 575 1300 • TWX 7-0 734 4314 • TELEX 6853128

# Appendix I. Program Listing . . . Con't.

6. Enter y-parameters or  $Y_s$  and  $Y_L$

Applicable Equations:

None

Flags:

03 set when y-parameters entered

04 set when  $Y_s$  and  $Y_L$  entered

01*LBL "PARA"	25 PROMPT	49 -
02 ADV	26 STO 05	50 STO 14
03 "ENTER Y-PARAMS"	27 "BO=?"	51 SF 03
04 AVIEW	28 PROMPT	52 RTN
05 ADV	29 STO 04	53*LBL "YS/L"
06 "GI=?"	30*LBL "PAR1"	54 ADV
07 PROMPT	31 RCL 06	55 "ENTER YS AND YL"
08 STO 03	32 RCL 07	56 AVIEW
09 "BI=?"	33 RCL 08	57 "GS=?"
10 PROMPT	34 RCL 09	58 PROMPT
11 STO 02	35 XEQ "Z"	59 STO 19
12 "GR=?"	36 STO 11	60 "BS=?"
13 PROMPT	37 X<>Y	61 PROMPT
14 STO 07	38 STO 10	62 STO 18
15 "BR=?"	39 X<>Y	63 "GL=?"
16 PROMPT	40 R-P	64 PROMPT
17 STO 06	41 STO 12	65 STO 21
18 "GF=?"	42 RCL 03	66 "BL=?"
19 PROMPT	43 RCL 05	67 PROMPT
20 STO 09	44 *	68 STO 20
21 "BF=?"	45 2	69 SF 04
22 PROMPT	46 *	70 ENG
23 STO 08	47 STO 13	
24 "GG=?"	48 RCL 11	

7. Calculate Stern Stability Factor or Specify Terminations for Given k

Applicable Equations:

$$P = R_e(y, y_i)$$

$$Q = I_m(y, y_i)$$

$$M = |y_i y_i|$$

To calculate k:

$$k = \frac{2g_i g_o (1 + \frac{G_s}{g_i} (1 + \frac{G_L}{g_o}))}{P + M}$$

To specify terminations:

if  $g_i > 0, g_o > 0$

then,

$$\frac{G_s}{|g_i|} = \frac{G_L}{|g_o|} = \sqrt{\frac{k(P + M)}{2|g_i g_o|}} - 1$$

if  $g_i < 0, g_o > 0$

or  $g_i > 0, g_o < 0$

then,

$$\frac{G_s}{|g_i|} = \frac{G_L}{|g_o|} = \sqrt{\frac{k(P + M)}{2|g_i g_o|}} - 1$$

if  $g_i < 0, g_o < 0$

then,

$$\frac{G_s}{|g_i|} = \frac{G_L}{|g_o|} = \sqrt{\frac{k(P + M)}{2|g_i g_o|}} + 1$$

## If We Don't Already Have The RF Filter You Need... We'll Build It, Fast.



Catalog RF/82 shows duplexers and bandpass, band reject and low pass/high pass filters currently being used by many manufacturers of transceivers and other VHF/UHF equipments.

But if you need a one-of-a-kind special and you can't afford to wait, we've still got you covered—we'll design and build exactly what you need for your system, and we'll work around the clock to deliver it when you need it.

Call us and talk to the RF engineer who will design your special filter. He'll give you a prompt, on-line analysis of your specifications, and he'll quote price and delivery time. Before you hang up, you'll know what you need, when you'll have it and how much it will cost—all with just one phone call!

Once you've placed an order, our unique QRC (quick reaction capability) begins to work for you: QRC combines the efficiency of computer-aided design with our dedicated model shop and test labs to ensure that your filter will be what you need when you need it.

When you need a special filter designed exactly to your specifications, and you need it *now*, call MFC!

**MFC**  
MICROWAVE FILTER COMPANY, INC.

315-437-3953  
TWX 710-541-0493  
6743 Kinne St., East Syracuse, NY 13057

# Flags:

When calculating k,  
07 set if either or both

$$(g_i + G_s) < 0$$

$$(g_o + G_L) < 0$$

When calculating terminations, 05 and 06 are set when  
the convergence is complete.

```

01*LBL "SPAR" 49 RCL 37 97 X<Y
02 "MS11=?" 50 XEQ "Z*" 98 STO 04
03 PROMPT 51 STO 28 99 -2
04 STO 31 52 X<Y 100 RCL 34
05 "AS11=?" 53 STO 27 101 *
06 PROMPT 54 1 102 -2
07 STO 30 55 RCL 31 103 RCL 35
08 "MS12=?" 56 + 104 *
09 PROMPT 57 STO 29 105 RCL 41
10 STO 35 58 RCL 30 106 RCL 42
11 "AS12=?" 59 X<Y 107 XEQ "Z/"
12 PROMPT 60 1 108 STO 07
13 STO 34 61 RCL 33 109 X<Y
14 "MS21=?" 62 + 110 STO 06
15 PROMPT 63 STO 38 111 -2
16 STO 37 64 RCL 32 112 RCL 36
17 "AS21=?" 65 X<Y 113 *
18 PROMPT 66 XEQ "Z*" 114 -2
19 STO 36 67 RCL 27 115 RCL 37
20 "MS22=?" 68 RCL 28 116 *
21 PROMPT 69 XEQ "Z-" 117 RCL 41
22 STO 33 70 STO 42 118 RCL 42
23 "AS22=?" 71 X<Y 119 XEQ "Z/"
24 PROMPT 72 STO 41 120 STO 09
25 STO 32 73 2 121 X<Y
26 29.037 74 RCL 29 122 STO 08
27 STO 26 75 - 123 2.009
28*LBL 01 76 STO 39 124 STO 26
29 ISG 26 77 2 125*LBL 04
30 GTO 02 78 RCL 38 126 .02
31 GTO 03 79 - 127 RCL IND
32*LBL 02 80 STO 40 128 *
33 RCL IND 26 81 RCL 30 129 STO IND
34 ISG 26 82 CHS 130 CHS
35 RCL IND 26 83 RCL 39 131 ISG 26
36 P-R 84 RCL 32 132 GTO 04
37 STO IND 26 85 RCL 38 133 RTN
38 1 86 XEQ 05 134*LBL 05
39 ST- 26 87 STO 03 135 XEQ "C*"
40 RDN 88 X<Y 136 RCL 27
41 X<Y 89 STO 02 137 RCL 28
42 STO IND 26 90 RCL 30 138 XEQ "Z+"
43 ISG 26 91 RCL 29 139 RCL 41
44 GTO 01 92 RCL 32 140 RCL 42
45*LBL 03 93 CHS 141 XEQ "Z/"
46 RCL 34 94 RCL 40 142 END
47 RCL 35 95 XEQ 05
48 RCL 36 96 STO 05

```

## 8. S-Parameter to y-Parameter Conversion

Applicable Equations:

$$y_i = \frac{(1 - S_{11})(1 + S_{22}) + S_{12}S_{21}}{(1 + S_{11})(1 + S_{22}) - S_{12}S_{21}}$$

$$y_o = \frac{(1 + S_{11})(1 - S_{22}) + S_{12}S_{21}}{(1 + S_{11})(1 + S_{22}) - S_{12}S_{21}}$$

$$y_r = \frac{-2S_{12}}{(1 + S_{11})(1 + S_{22}) - S_{12}S_{21}}$$

$$y_i = \frac{-2S_{21}}{(1 + S_{11})(1 + S_{22}) - S_{12}S_{21}}$$

$$y = y' \left( \frac{1}{z_o} \right),$$

where  $z_o$  is the S-parameter reference impedance,  
 $1/z_o$  is set to 0.02 in the program (step 126).

Flags:

03 set when y-parameters are calculated.

```

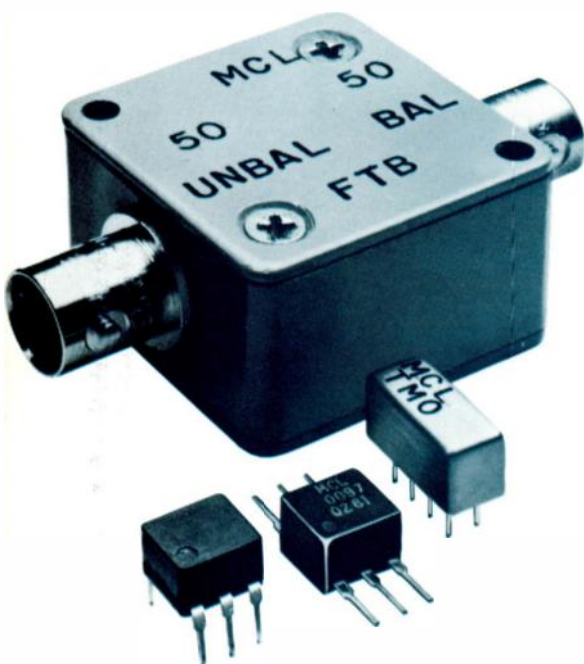
01*LBL "STM" 40 ABS 79 CF 06
02 CF 07 41 / 80 RCL 02
03 RCL 03 42 STO 27 81 STO 22
04 RCL 19 43 RCL 03 82 RCL 04
05 + 44 RCL 05 83 STO 24
06 X<0? 45 * 84*LBL 01
07 SF 07 46 X<0? 85 RCL 22
08 RCL 05 47 GTO 02 86 CHS
09 RCL 21 48 RCL 03 87 STO 18
10 + 49 X<0? 88 RCL 24
11 X<0? 50 GTO 03 89 CHS
12 SF 07 51 RCL 27 90 STO 20
13 * 52 SQRT 91 XEQ "Y1/O"
14 2 53 1 92 RCL 22
15 * 54 - 93 RCL 18
16 RCL 11 55 GTO 05 94 +
17 RCL 12 56*LBL 02 95 RCL 22
18 + 57 RCL 27 96 /
19 / 58 1 97 "ABS"
20 STO 17 59 + 98 1 E-4
21 FC? 07 60 SQRT 99 X<Y?
22 RTN 61 GTO 05 100 SF 05
23 ABS 62*LBL 03 101 RCL 24
24 CHS 63 RCL 27 102 RCL 20
25 STO 17 64 SQRT 103 +
26 RTN 65 1 104 RCL 24
27*LBL "ITR" 66 + 105 /
28 "K=?" 67*LBL 05 106 ABS
29 PROMPT 68 STO 26 107 1 E-4
30 STO 17 69 RCL 03 108 X<Y?
31 RCL 11 70 ABS 109 SF 06
32 RCL 12 71 * 110 FC? 05
33 + 72 STO 19 111 GTO 01
34 * 73 RCL 26 112 FC? 06
35 2 74 RCL 05 113 GTO 01
36 / 75 ABS 114 SF 04
37 RCL 03 76 * 115 BEEP
38 RCL 05 77 STO 21 116 END
39 * 78 CF 05

```



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## DC ISOLATED PRIMARY & SECONDARY



Model No.	T1-1	T1-1H	T1.5-1	T2.5-6	T4-6	T9-1	T9-1H	T16-1	T16-1H
Imped. Ratio	1	1	1.5	2.5	4	9	9	16	16
Freq. (MHz)	.15-400	8-300	.1-300	.01-100	.02-200	.15-200	2-90	.3-120	7-85
T Model (10-49)	\$2.95	\$4.95	\$3.95	\$3.95	\$3.95	\$3.45	\$5.45	\$3.95	\$5.95
TMO model (10-49)	\$4.95		\$6.75	\$6.45	\$6.45	\$6.45		\$6.45	

## CENTER-TAPPED DC ISOLATED PRIMARY & SECONDARY

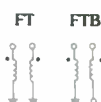


Model No.	T1-1T	T2-1T	T2.5-6T	T3-1T	T4-1	T4-1H	T5-1T	T13-1T
Imped. Ratio	1	2	2.5	3	4	4	5	13
Freq. (MHz)	.05-200	.07-200	.01-100	.05-250	2-350	8-350	3-300	.3-120
T Model (10-49)	\$3.95	\$4.25	\$4.25	\$3.95	\$2.95	\$4.95	\$4.25	\$4.25
TMO model (10-49)	\$6.45	\$6.75	\$6.75	\$6.45	\$4.95		\$6.75	\$6.75

## UNBALANCED PRIMARY & SECONDARY



Model No.	T2-1	T3-1	T4-2	T8-1	T14-1
Imped. Ratio	2	3	4	8	14
Freq. (MHz)	.025-600	5-800	2-600	.15-250	2-150
T model (10-49)	\$3.45	\$4.25	\$3.45	\$3.45	\$4.25
TMO Model (10-49)	\$5.95	\$6.95	\$5.95	\$5.95	\$6.75



Model No.	FT1.5-1	FTB1-1	FTB1-6	FTB1-1-75
Imped. Ratio	1.5	1	1	1
Freq. (MHz)	.1-400	.2-500	.01-200	.5-500
(1-4)	\$29.95	\$29.95	\$29.95	\$29.95

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INFO/CARD 40

WRN

59 Rev. B

## Appendix I. Program Listing . . . Con't.

### 9. Complex Arithmetic

Routines do complex addition, subtraction, multiplication, division and absolute value. Most are similar to those in the HP Math Pac I module.

Applicable Equations:

Standard complex arithmetic.

Flags:

None

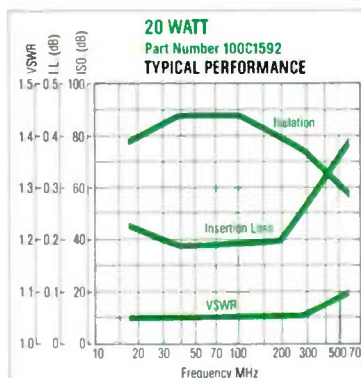
01 LBL "Z+"	16 LBL 02	35 *
02 X<>Y	19 R-P	36 RDN
03 LBL 05	20 1/X	37 +
04 RDN	21 X<>Y	38 R↑
05 +	22 CHS	39 P-R
06 RDN	23 X<>Y	40 RTN
07 +	24 P-R	41 LBL "Z/"
08 R↑	25 RTN	42 XEQ 02
09 RTN	26 LBL "Z*"	43 GTO "Z*"
10 LBL "Z-"	27 R-P	44 LBL "ZABS1"
11 CHS	28 RDN	45 R-P
12 X<>Y	29 RDN	46 RDN
13 CHS	30 R-P	47 RDN
14 GTO 05	31 RDN	48 RCL Z
15 LBL "ZINV"	32 RDN	49 END
16 XEQ 02	33 X<>Y	
17 RTN	34 RDN	

## Appendix II. Storage Assignments.

Register	Quantity	Register	Quantity	Reg.	Qty.
00	Not used	18	$B_S$	36	$\text{Im}(S_{21})$
01	Not used	19	$G_S$	37	$\text{Re}(S_{21})$
02	$b_i$	20	$B_L$	38	scratch
03	$g_i$	21	$G_L$	39	scratch
04	$b_o$	22	$B_{in}$	40	scratch
05	$g_o$	23	$G_{in}$	41	scratch
06	$b_r$	24	$B_{out}$	42	scratch
07	$g_r$	25	$G_{out}$		
08	$b_f$	26	scratch	registers	
09	$g_f$	27	scratch	30-42 used	
10	$Q = \text{Im}(Y_r y_i)$	28	scratch	only by S-	
11	$P = \text{Re}(y_r y_i)$	29	scratch	parameter	
12	$M =  y_r y_i $	30	$\text{Im}(S_{11})$	routine	
13	$2g_r g_o$	31	$\text{Re}(S_{11})$		
14	$2g_r h_o - P$	32	$\text{Im}(S_{22})$		
15	$(2g_r g_o - P)^2 - M^2$	33	$\text{Re}(S_{22})$		
16	C(Linville factor)	34	$\text{Im}(S_{12})$		
17	k (Stern factor)	35	$\text{Re}(S_{12})$		

# 20 & 100 WATT

## RF SOLID-STATE SWITCHES



### Specifications

20 WATT P/N 100C1592

Configuration: SP2T

Frequency: 20-500 MHz

RF Power: 20 WATT CW

Control: 1 Line TTL

Switching

Speed: 10  $\mu$ sec. Max.

DC Power: +5V at 220 mA

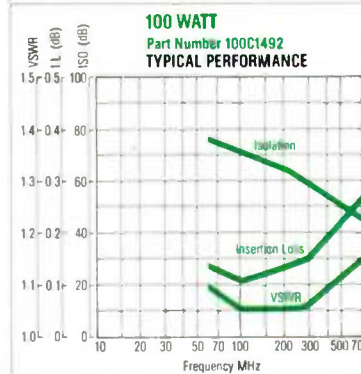
-15V to -30V

at 20 mA

Impedance: 50 Ohms

Connectors: SMA

Size: 2.75" x 3" x 1.4"



### Specifications

100 WATT P/N 100C1492

Configuration: SP2T

Frequency: 100-500 MHz

RF Power: 100 WATT CW

Control: 1 Line TTL

Switching

Speed: 30  $\mu$ sec. Max.

DC Power: +5V at 300 mA

-50V at 10 mA

Impedance: 50 Ohm

Connectors: N

Size: 4" x 4.75" x 1.3"



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INFO/CARD 41

# Lumped-Constant Line Stretcher For Testing Power Amplifier Stability

To determine the stability of an RF power amplifier, oscillator, or transmitter under harsh conditions, a variable line and fixed attenuator are used to simulate a given VSWR at all phase angles. This article describes a lumped-constant line stretcher using either variable capacitors or inductors.

By E.A. Franke and A.E. Noorani  
General Electric Company  
Lynchburg, VA 24502

**T**ransmitters employing electron tubes were typically narrowband tuned devices. With the advent of the transistor, the RF power amplifier is capable of wideband operation and must therefore be evaluated for stability to avoid interference with other signals. The spectral purity must be maintained in spite of variations in temperature, supply voltage, or load impedance. For applications such as marine, automotive, or portable operation, the antenna may present a severe mismatch to the power amplifier due to the proximity of foreign objects, environmental corrosion, or a pinched coaxial cable. A most stringent test of stability occurs when the power amplifier operates into a high-Q load such as a narrowband duplexer or cavity filter. Commercial mobile transmitters are typically rated to maintain stability into a 3:1 VSWR even under fluctuations of  $\pm 20\%$  in supply voltage.

The power amplifier may be tested into a worse case mismatch at all phase angles. This is easily simulated by connecting the amplifier output to an attenuator followed by an adjustable air line terminated in either an open or a short, (Figure 1). Stability is monitored on a spectrum analyzer connected to a directional coupler placed between the power amplifier and the attenuator. Movement of the air line through one-half wavelength will traverse a complete circle around the Smith Chart. The radius of the circle will depend on the size of attenuator used. The mismatch presented by the artificial load is best described using return loss. The value of return loss is simply twice the attenuation

of the fixed pad, assuming negligible loss in the adjustable air line. The mismatch in terms of standing wave ratio is given as

$$\text{VSWR} = \coth \left[ \frac{\text{attenuator (dB)}}{8.686} \right] \quad (1)$$

The radius of the mismatch circle on the normalized Smith Chart intersects the right half of the abscissa at the value of the VSWR.

Recently Roderick Blocksom<sup>1</sup> reported on a binary stepped transmission line formed by numerous lengths of coaxial cable and relays. Below 200 MHz the length of a half-wave adjustable air line ( $>75$  cm) becomes so cumbersome that a continuously-variable, lumped-constant version is needed. This article suggests a stretcher using either variable capacitors or inductors instead of transmission lines.

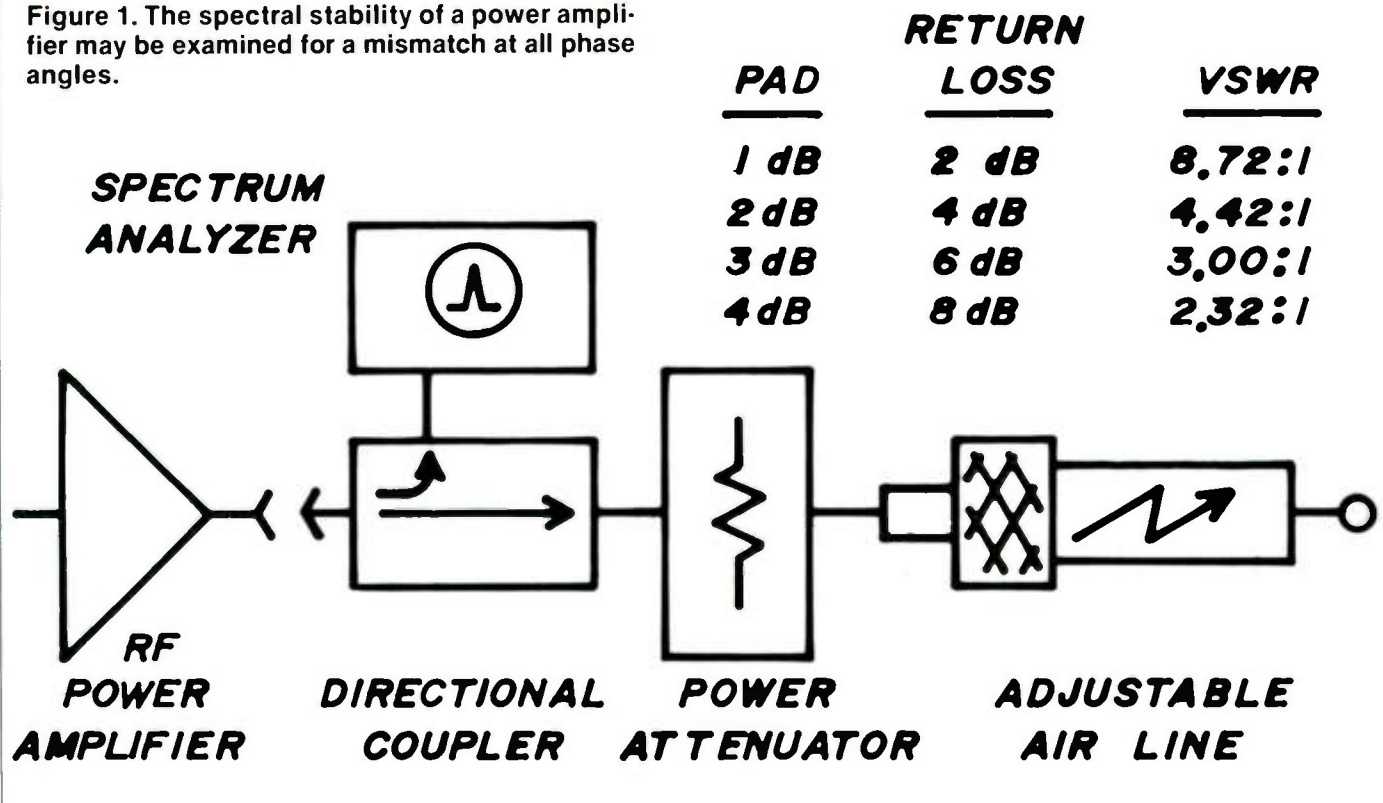
## Lumped-Constant Line Stretcher

Consider the network shown in Figure 2. The input impedance is simply the parallel combination of two series circuits  $L_1, C_1$  and  $L_2, C_2$ .

$$Z_{in} = j \frac{\left( \omega L_1 - \frac{1}{\omega C_1} \right) \left( \omega L_2 - \frac{1}{\omega C_2} \right)}{\omega(L_1 + L_2) - \frac{1}{\omega} \left( \frac{1}{C_1} + \frac{1}{C_2} \right)} \quad (2)$$

where  $\omega = 2\pi f$ .

Figure 1. The spectral stability of a power amplifier may be examined for a mismatch at all phase angles.



Examination of the above equation shows two zeros in the numerator corresponding to two series resonances. A parallel resonance occurs in the denominator when

$$\omega(L_1 + L_2) = \frac{2}{\omega C} \quad (3)$$

where  $C_1 = C_2 = C$  and assuming a ganged dual section or butterfly variable capacitor. The above equation may be rewritten as

$$XC = \frac{XL_1 + XL_2}{2}$$

where  $XL_1 = \omega L_1$ ,  $XL_2 = \omega L_2$  and  $X_C = \frac{1}{\omega C}$ .

At any given frequency, the locus of impedance points plotted on a Smith Chart as the capacitor is varied will be a circle similar to that achieved using a variable length transmission line. If an attenuator is not used, the points will be entirely imaginary ( $R = 0$ ). If an attenuator is used, the value of the real and imaginary parts of the input impedance may simply be read from the chart. When properly designed, as the capacitance is increased, the locus travels from a short circuit (when  $XL_1 = X_C$ ) (Figure 3) to an inductive

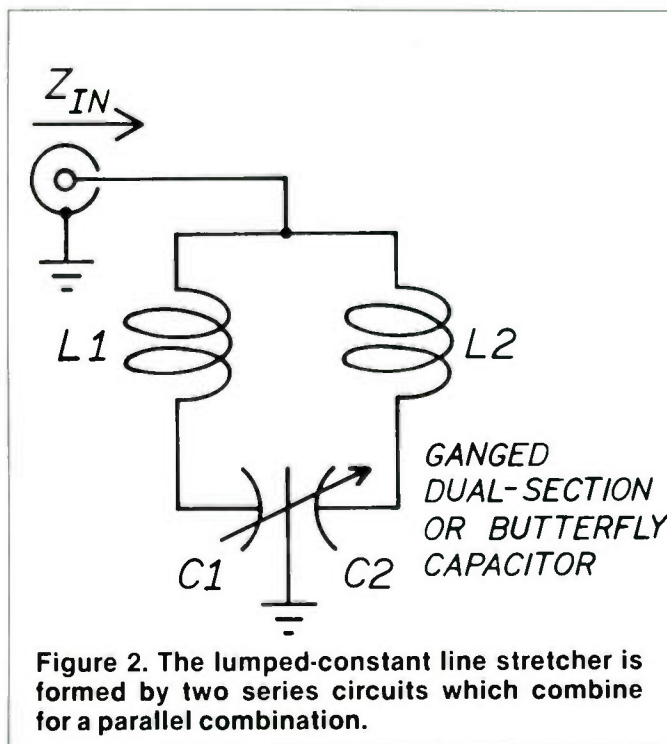


Figure 2. The lumped-constant line stretcher is formed by two series circuits which combine for a parallel combination.

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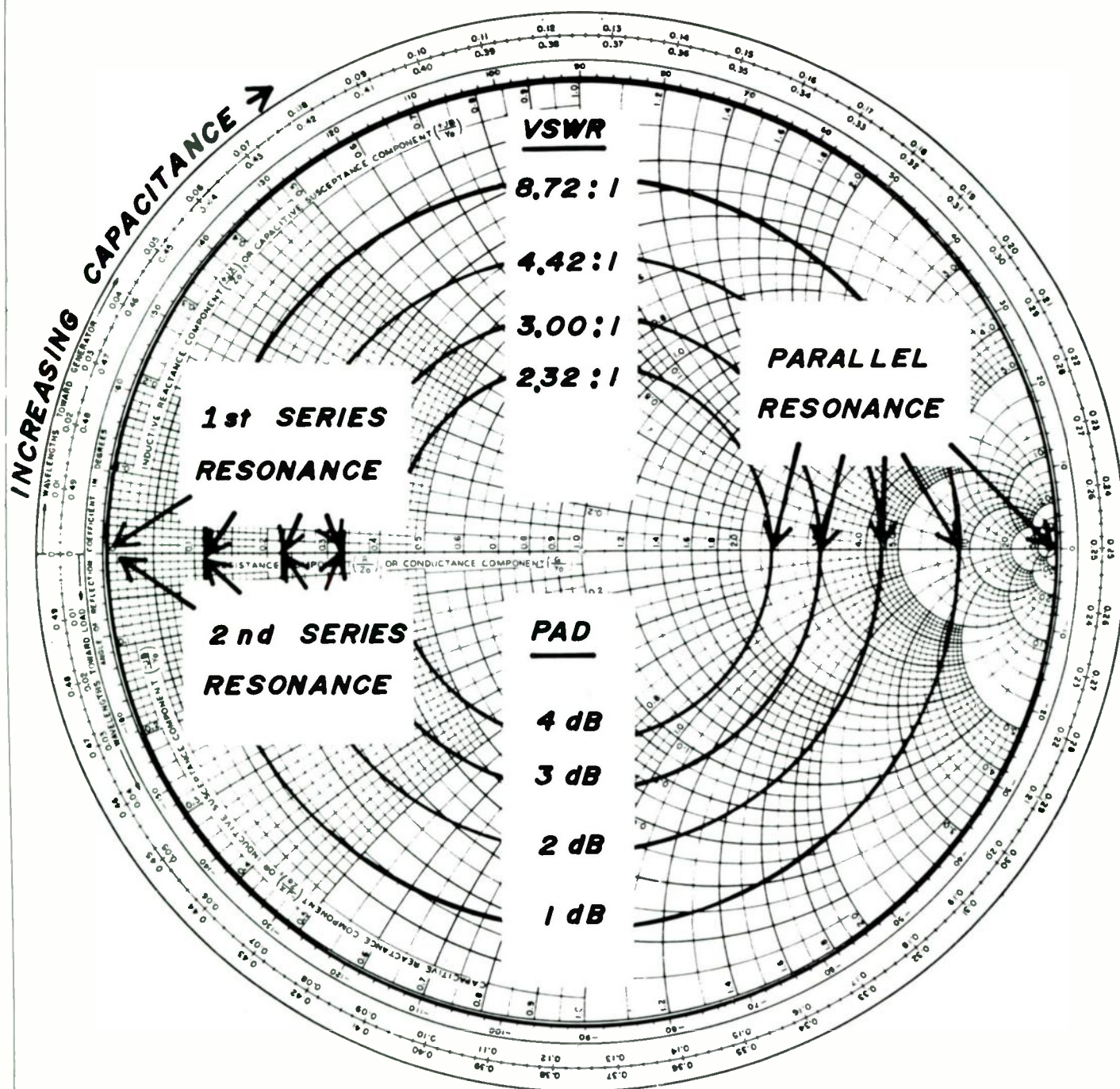


Figure 3. As the variable capacitor is meshed, the loci of input impedance points forms a clockwise circle of constant VSWR covering all phase angles.

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reactance, then through parallel resonance to a capacitive reactance. Finally the locus passes through the second series resonance (when  $XL_2 = X_C$ ) near maximum capacitance.

By properly dictating the end points of the locii at the series resonances, the parallel resonance will occur somewhere between them.

For design purposes, it is necessary to establish the frequency bandwidth over which the lumped-constant variable line has to work. Next the ganged dual-section or butterfly capacitor selected must have a high breakdown voltage and a very high Q. Any dissipative loss in the coil or capacitor will be evidenced by a non-circular response on the Smith Chart. Before any calculations for the inductors are made, the minimum and maximum capacitance should be measured. It is also necessary that the capacitor does not have a self resonance anywhere in the frequency band of interest.

Inductor  $L_1$  is calculated using the maximum capacitance at the lowest operating frequency for series resonance.

$$L_1 = \frac{1}{4\pi^2(F_{\text{low}})^2 C_{\text{max}}} \quad (4)$$

The second inductor  $L_2$  is calculated using the minimum variable capacitance at the highest operating frequency.

$$L_2 = \frac{1}{4\pi^2(F_{\text{high}})^2 C_{\text{min}}} \quad (5)$$

The ratio of inductance  $L_1/L_2$  must be less than unity

$$\frac{L_1}{L_2} = \left( \frac{F_{\text{high}}}{F_{\text{low}}} \right)^2 \left( \frac{C_{\text{min}}}{C_{\text{max}}} \right), \quad (6)$$

which reduces to

$$\left( \frac{F_{\text{high}}}{F_{\text{low}}} \right)^2 < \left( \frac{C_{\text{max}}}{C_{\text{min}}} \right) \quad (7)$$

If the inductors are equal, then parallel resonance will occur at the same time as series resonance both at the highest frequency ( $C_{\text{min}}$ ) and at the lowest frequency ( $C_{\text{max}}$ ).

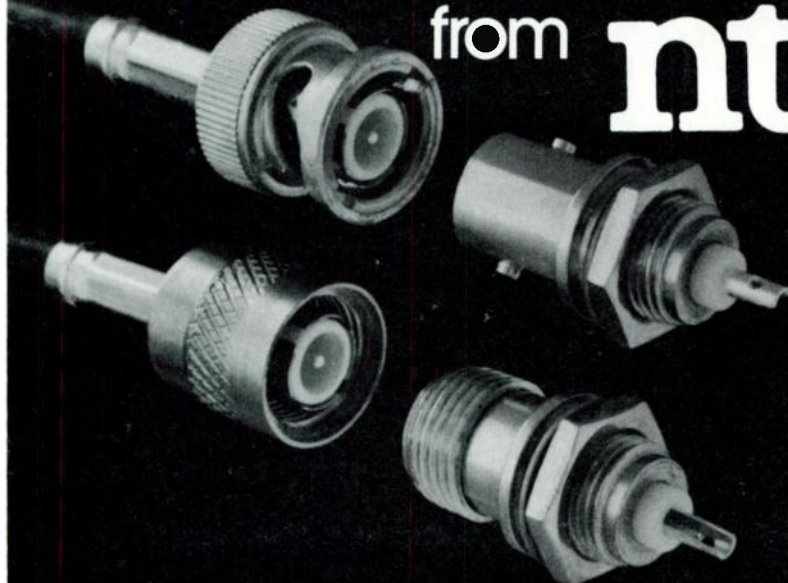
If the ratio of  $L_1/L_2$  is greater than unity, parallel resonance will not occur within the frequency range  $F_{\text{low}}$  to  $F_{\text{high}}$ .

From equation 7, the operating range is limited by the square root of the capacitance ratio. It should be remembered that the input impedance will vary faster when a wider frequency range is chosen. Thus to insure a more gradual change in impedance as the capacitor is rotated, the frequency range is usually limited only to the band of interest.

## VHF Example

As an example calculation, a variable line was designed to operate over the two-way radio band, 150 to 175 MHz. A dual variable capacitor was chosen with a minimum capacitance of 7.5 pF and a maximum of 50 pF. We shall limit the range of capacitance to  $C_{\text{min}} = 10$  pF and  $C_{\text{max}} = 45$  pF to account for parasitic capacitance and to operate slightly away from the capacitor stops.

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$$L_1 = \frac{1}{4 \pi^2 (F_{low})^2 C_{max}} = \frac{1}{4 \pi^2 (150 \text{ MHz})^2 (45 \text{ pF})} = 25 \text{ nH}$$

$$L_2 = \frac{1}{4 \pi^2 (F_{high})^2 C_{min}} = \frac{1}{4 \pi^2 (175 \text{ MHz})^2 (10 \text{ pF})} = 83 \text{ nH}$$

Very low values of inductance may be simply short lengths of wire or flat copper strips. The exact value is determined by adjusting for series resonance at each end of the capacitor travel.

The input impedance may be quickly examined at band edges. At 150 MHz the first series resonance occurs at  $C_{max} = 45 \text{ pF}$  by the above design. At the second series resonance, the variable capacitor resonates with the other inductor  $L_2$ .

$$XC = XL_2 = 2\pi f L_2 = 2\pi (150 \text{ MHz})(83 \text{ nH}) = 78 \text{ ohms}$$

$$C = \frac{1}{2\pi f X_C} = \frac{1}{2\pi (150 \text{ MHz})(78 \text{ ohm})} = 13.6 \text{ pF}$$

A parallel resonance will occur between the two series resonances when all of the components are taken into account.

$$XC = \frac{XL_1 + XL_2}{2} = \frac{2\pi f (L_1 + L_2)}{2}$$

$$= \pi (150 \text{ MHz})(25 \text{ nH} + 83 \text{ nH}) = 50.9 \text{ ohm}$$

$$C = \frac{1}{2\pi f XC} = \frac{1}{2\pi (150 \text{ MHz})(78 \text{ ohm})} = 20.8 \text{ pF}$$

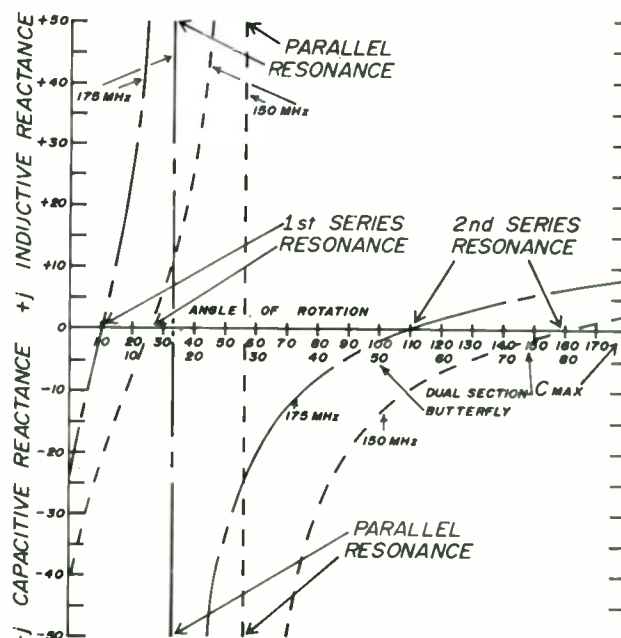


Figure 4.

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Model Number (2)	Impedance Ohms (Power W)	Frequency Range	BNC	TNC	N	SMA	UHF	PC
<b>Fixed Attenuators, 1 to 20 dB</b>								
AT-50(3)	50 (5W)	DC-1.5GHz	14.00	20.00	20.00	18.00	—	—
AT-51	50 (5W)	DC-1.5GHz	11.00	15.00	15.00	14.00	—	—
AT-52	50 (1W)	DC-1.5GHz	14.50	20.50	20.50	19.50	—	12.00
AT-53	50 (25W)	DC-3.0GHz	14.00	17.00	—	15.00	—	—
AT-54	50 (25W)	DC-4.2GHz	—	—	—	18.00	—	—
AT-75 or 93	75 or 93 (5W)	DC-1.5GHz (750MHz)	14.00	20.00	20.00	18.00	—	—
<b>Detector, Zero Bias Schottky</b>								
CD-51	50	0.1-4.2GHz	—	—	—	54.00	—	—
<b>Resistive Impedance Transformers, Minimum Loss Pads</b>								
RT-50 75	50 to 75	DC-1.5GHz	10.50	19.50	19.50	17.50	—	—
RT-50 93	50 to 93	DC-1.0GHz	13.00	19.50	19.50	17.50	—	—
<b>Terminations</b>								
CT-50 (3)	50 (5W)	DC-4.2GHz	11.50	15.00	15.00	17.50	—	—
CT-51	50 (5W)	DC-4.2GHz	9.50	12.00	12.00	9.50	—	—
CT-52	50 (1W)	DC-2.5GHz	10.50	15.00	15.00	13.00	15.50	—
CT-53 M	50 (5W)	DC-4.2GHz	—	—	—	5.60 (10 Pcs)	—	—
CT-54	50 (2W)	DC-2.0GHz	14.00	15.00	15.00	17.50	—	—
CT-75	75 (25W)	DC-2.5GHz	10.50	15.00	15.00	13.00	15.50	—
CT-93	93 (25W)	DC-2.5GHz	13.00	15.00	—	—	15.50	—
<b>Mismatched Terminations, 105:1 to 3:1 Open Circuit, Short Circuit</b>								
MT-51	50	DC-3.0GHz	29.50	29.50	—	25.50	—	—
MT-75	75	DC-1.0GHz	—	25.50	—	—	—	—
<b>Feed thru Terminations, shunt resistor</b>								
FT-50	50	DC-1.0GHz	10.50	19.50	19.50	17.50	—	—
FT-75	75	DC-500MHz	10.50	19.50	19.50	17.50	—	—
FT-93	93	DC-150MHz	13.00	19.50	19.50	17.50	—	—
<b>Directional Coupler, 30 dB</b>								
DC-500	50	250-500MHz	80.00	—	—	—	—	—
<b>Resistive Decoupler, series resistor or Capacitive Coupler, series capacitor</b>								
RD or CC-1000	1000 (1000PF)	DC-1.5GHz	12.00	18.00	18.00	17.00	—	—
<b>Adapters</b>								
CA-50 (N to SMA)	50	DC-4.2GHz	—	—	13.00	13.00	—	—
<b>Inductive Decouplers, series inductor</b>								
LD-R15	0.17uH	DC-500MHz	12.00	18.00	18.00	17.00	—	—
LD-R80	6.8uH	DC-55MHz	12.00	18.00	18.00	17.00	—	—
<b>Fixed Attenuator Sets, 3, 6, 10, and 20 dB, in plastic case</b>								
AT-50-SET (3)	50	DC-1.5GHz	60.00	84.00	84.00	76.00	—	—
AT-51-SET	50	DC-1.5GHz	48.00	84.00	84.00	80.00	—	—
<b>Reactive Multicouplers, 2 and 4 output ports</b>								
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RC-3-30	50	DC-500MHz	54.00	—	—	54.00	—	—
RC-9-30	50	DC-500MHz	—	—	—	84.50	—	—
<b>Double Balanced Mixers</b>								
DBM-1000	50	5-1000MHz	61.00	—	71.00	61.00	—	—
DBM-500PC	50	2-500MHz	—	—	—	—	34.00	—
<b>RF Fuse, 1.8 Amp, and 1.6 Amp</b>								
FL-50	50	DC-1.5 GHz	12.00	18.00	—	17.00	—	—
FL-75	75	DC-1.5 GHz	12.00	18.00	—	17.00	—	—

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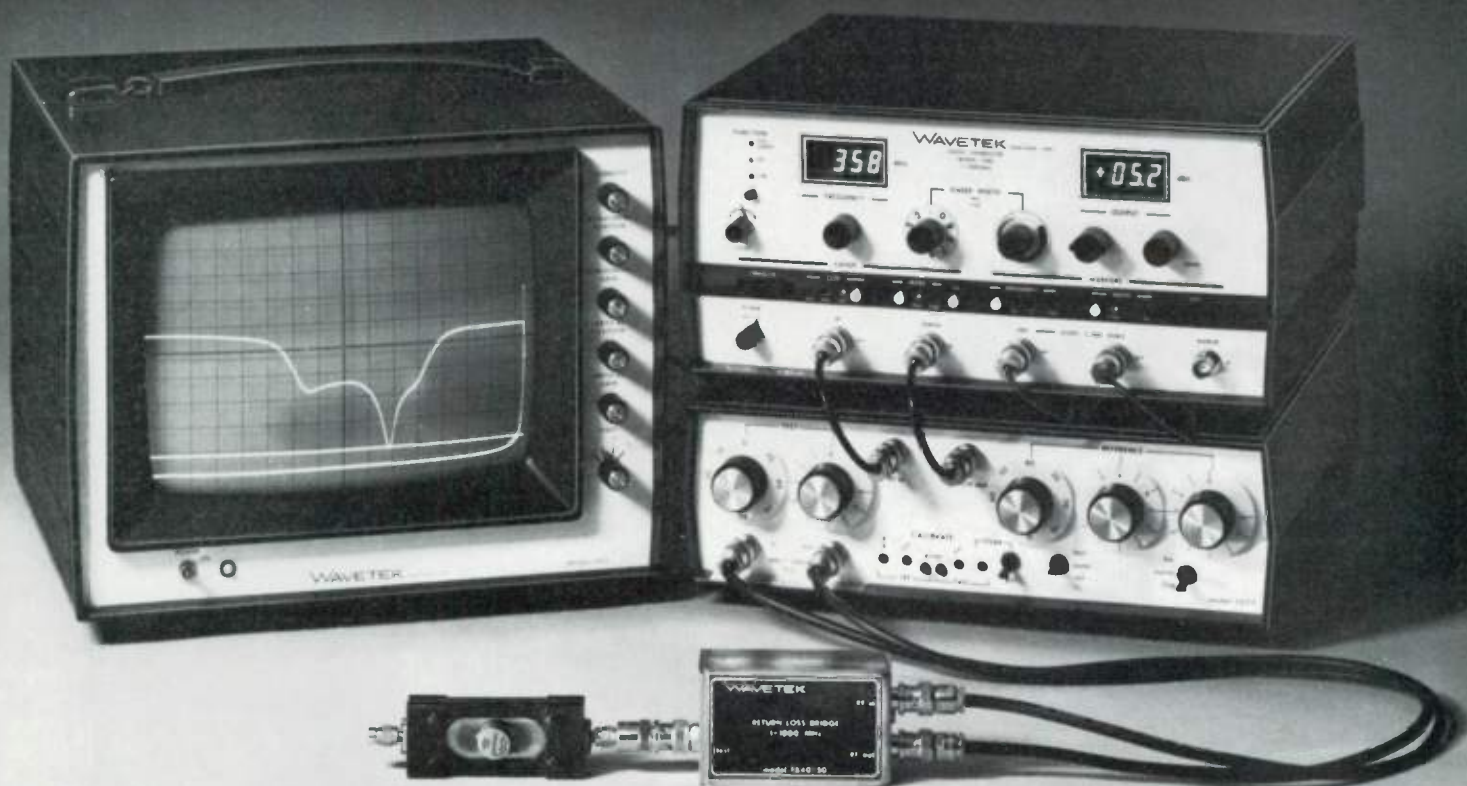
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Demonstration: INFO/CARD 45 Literature: INFO/CARD 73

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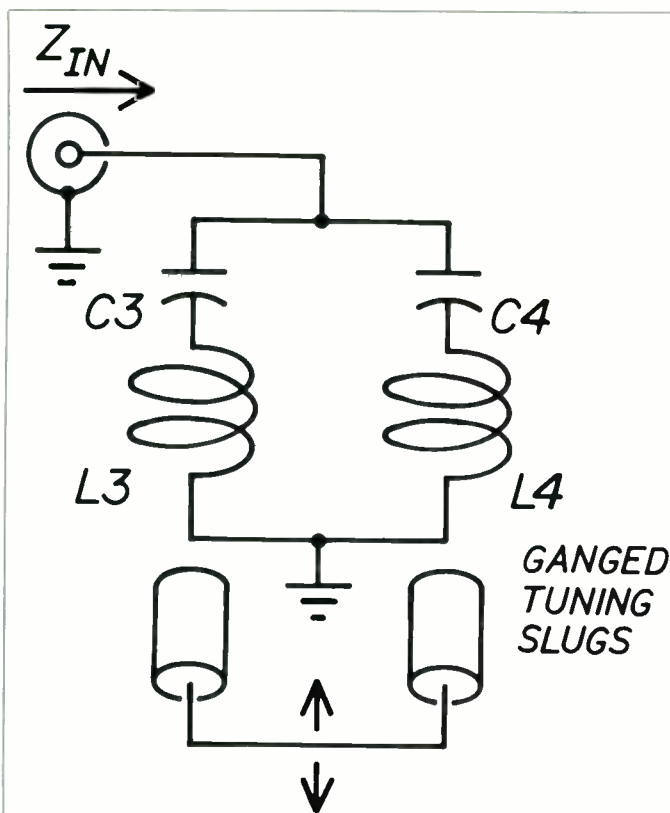


Figure 5. The lumped-constant adjustable line simulator may also be formed using a permeability tuner.

At the high end of the band (175 MHz) the following conditions are present:

First Series Resonance  $C = 33 \text{ pF}$   
 Parallel Resonance  $C = 15.3 \text{ pF}$   
 Second Resonance  $C = 10 \text{ pF}$

The input impedance is plotted as a function of capacitor rotation for the lowest and highest operating frequency in Figure 4. Rotation is shown both for a semi-circular ( $180^\circ$ ) and a butterfly ( $90^\circ$ ) capacitor. A gear reduction is typically employed with a calibrated dial. The equation for the capacitor is assumed to be linear

$$C = [(\theta/180^\circ)(C_{\max} - C_{\min})] + C_{\min}$$

where

$\theta$  = angular rotation in degrees from fully open

$C_{\max} = 50 \text{ pF}$

$C_{\min} = 7.5 \text{ pF}$

The allowable RF power input to this type of device is dependent on the breakdown voltage of the capacitor. For the above design a 500V breakdown capacitor was used and 150 Watts of RF power was applied without any adverse effects.

## Variable Inductance Line Stretcher

The variable may be switched from the capacitor to the inductor (Figure 5). A slug formed from ferrite, brass, copper, or aluminum may be mechanically positioned within two identical inductors to vary the inductance. The fixed capacitors, C3 and C4, are computed in a similar manner as above. The conductive metallic core acts as a single shorted turn to decrease the inductance with increased penetration. The powdered iron or ferrite core has the opposite effect

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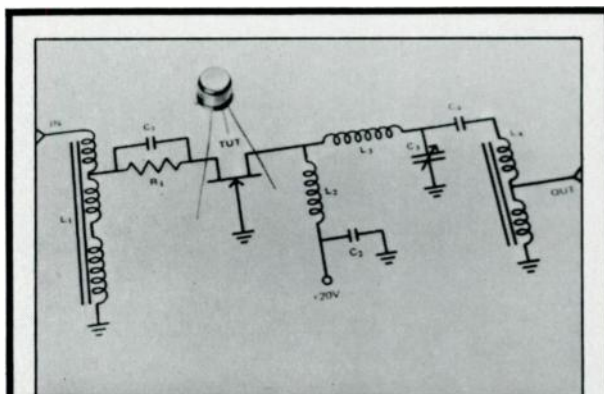
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and increases the inductance as it penetrates the coil. The Q of the coil is not appreciably lowered when using the brass slug provided the slug has a clean surface or is silver plated. The use of the powdered iron or ferrite core will raise the Q of the coil provided the material has been selected for the frequency range in use. Ferrite and powdered iron are typically usable up to 200 MHz.

## Variable Inductor Example

When a slug having a permeability of  $\mu$  is inserted into an air-core inductance  $L_{air}$ , the final inductance becomes  $\mu L_{air} = L_{core}$ . Thus  $L_{core}$  and  $L_{air}$  are the maximum and minimum values of inductance.

If the minimum and maximum of the matched coils are chosen for example to be  $L_{max} = L_{core} = 83 \text{ nH}$  and  $L_{min} = L_{air} = 25 \text{ nH}$ , the values of capacitance may be then calculated.

$$\mu = \frac{L_{core}}{L_{air}} = \frac{83 \text{ nH}}{25 \text{ nH}} = 3.3, \text{ ferrite}$$

$$C_3 = \frac{1}{4\pi^2 (F_{low})^2 L_{max}}$$

$$C_4 = \frac{1}{4\pi^2 (F_{high})^2 L_{min}}$$

For the frequency range of 150 to 174 MHz,  $C_3 = 13.56 \text{ pF}$  and  $C_4 = 33.08 \text{ pF}$ . Parallel resonance will occur when

$$2\omega L = \frac{1}{\omega} \left( \frac{1}{C_3} + \frac{1}{C_4} \right)$$

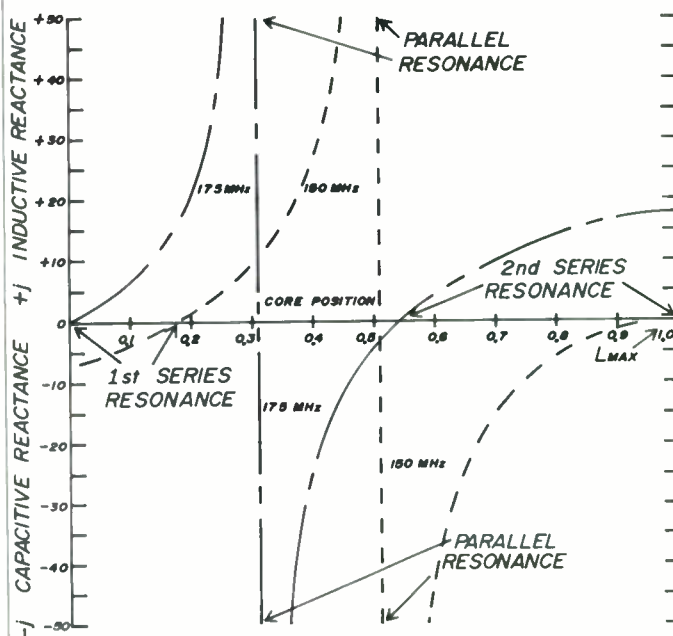


Figure 6. The input for the variable inductor is similar to the previous case as the ferrite slug is inserted into the coil.

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PF804	215-320	27	4.0	+35
PF7410C	406-512	16.5	4.5	+35
PF797A	800-960	19.5	5.0	+35

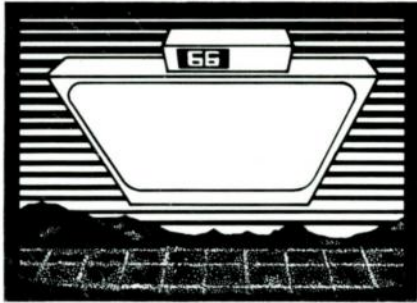
In addition to RF Amplifiers, Janel manufactures a wide range of standard Power Dividers and other rf components. Custom designs can be provided for unusual applications. For detailed information, call or write Janel Laboratories, Inc., 33890 Eastgate Circle, Corvallis, OR 97333. Telephone (503) 757-1134.



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$$L = \frac{C_3 + C_4}{2\omega^2 C_3 C_4} = 58.5 \text{ nH}$$

for parallel resonance. The value of each inductor will be 43 nH for parallel resonance at 175 MHz. This is plotted in Figure 6. The abscissa is plotted as linear travel for the position of a ferrite slug into a cylindrical, linearly-wound coil whose dimensions equal those of the core being used. It represents the fraction of the core inside the coil.

## Conclusion

To determine the stability of an RF power amplifier, oscillator, or transmitter under harsh conditions, a variable line and fixed attenuator are used to simulate a given VSWR at all phase angles. Below 200 MHz a half-wave adjustable line is cumbersome. This article has described a lumped-constant line stretcher using either variable capacitors or inductors.

The use of moveable slugs eliminates mechanical wear at the electrical grounding point present on the capacitor rotor. For automated testing of production transistors or amplifiers, a frictionless device is necessary.

Today there are few manufacturers still fabricating dual section or butterfly variable capacitors because their use was primarily in electron tube amplifiers.

## Reference

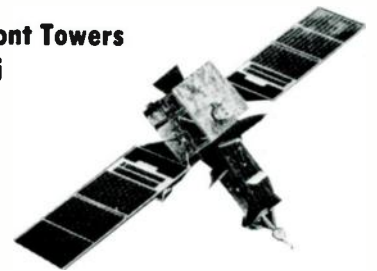
1. R.K. Blocksom, "A Binary Stepped Transmission Line," *r.f. design*, July/August 1982, pp. 22-29. □

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# An Easy Method for Measuring Unloaded Q ( $Q_u$ ) for an Inductor or Filter Tank

A measurement technique to measure inductance by using a narrowband single pole bandpass filter whose Q approximately equals the unloaded Q of the inductor or tank being measured.

By Marvin Kefer  
Hazeltine Corp.  
Greenlawn, N.Y.

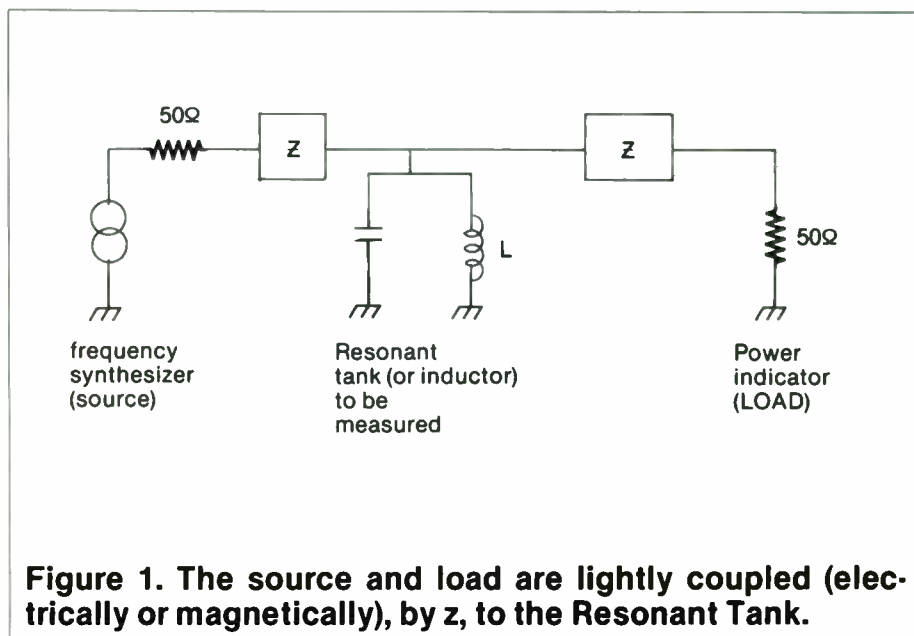
**T**his article presents an easy, accurate method for selecting the best (highest  $Q_u$ ) inductor or tank before you build a filter. Measurements are done at the operating frequency and include self resonant effects. The method of measurement uses readily available laboratory equipment, and is usable at any frequency.

The described technique uses a measurement circuit which is a narrow-band, single-pole, bandpass filter whose inductance (or tank) will be measured. Conceptually, the filter Q is dependent on (and limited by) the unloaded Q of the tank or inductor. By measuring the filter's Q, the tank or inductor is defined. The method is illustrated by an example which calculates the inductance and then resolves the inductor, or tank Q, from the filter's Q.

At low frequencies the coupling can be capacitive or inductive; at higher frequencies magnetic or electric probes are suggested.

A 285 MHz bandpass filter design requires a 75nH inductor. The measure-

ment of inductance and  $Q_u$  uses the test circuit shown in Figure 2. Light coupling and resonant capacitors are implemented by using accurately measured high Q series chip capacitors. It is tacitly presumed that  $Q_u$  (coupling



**Figure 1. The source and load are lightly coupled (electrically or magnetically), by z, to the Resonant Tank.**

## Construction of the Measurement Circuit

This measurement method requires that the resonated inductor, or tank, be lightly coupled to the source and load as shown in Figure 1. At low fre-

capacitor) >  $Q_u$  (resonant capacitor) >  $Q_u$  (inductor) and therefore the effects of these very high capacitor  $Q_u$ 's will be ignored: The parallel resonance equivalent circuit which is used to illustrate the calculations is shown in Figure 2. The series to parallel transformation (derived Appendix I) is used to convert the test circuit to the parallel resonant circuit in Appendix II.

Measurement of the 3dB bandpass frequencies is used to find  $Q$  and  $L$  as shown below (see Figure 3).

1.  $f_o = \sqrt{f_1 f_2} \approx (f_1 + f_2)/2$  (narrow-band approximation)  
 where  $f_1$  = lower 3dB frequency (283.5 MHz)  
 $f_2$  = higher 3dB frequency (286.4 MHz)  
 $= 284.95$  MHz
2.  $Q = f_o/BW(3dB)$   
 where  $BW(3dB) = f_2 - f_1 = 2.9$  MHz  
 $= 98.26$
3.  $L = 1/(2\pi f_o)^2 C$   
 where  $C = (3.96 + 0.1 + 0.1)$  pF (from Figure 2)  
 $= 75$  nH

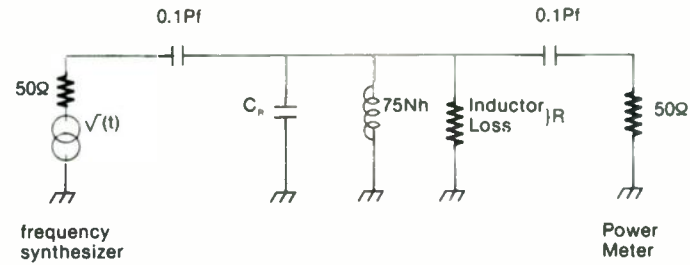
It should be noted that synthesizer frequency is usually very accurately displayed and measured. Overall measurement accuracy is dependent on the accuracy of  $C$  (capacitance) measurement and of the 3dB insertion loss measurement.

Figure 3 shows that the 3dB frequencies have well defined insertion loss. Therefore, they can be used to calculate  $f_o$  instead of directly measuring  $f_o$ , whose insertion loss is not well defined for small variations of  $f_o$  (Passband frequencies).

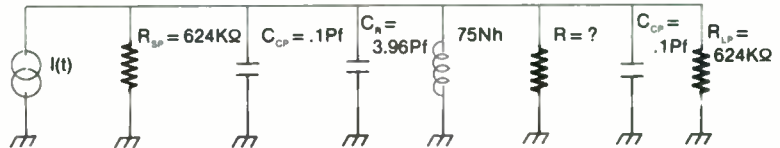
The  $Q$  calculated is equal to  $Q_u$  of the inductor or tank when the insertion loss of the test circuit is high as derived in Appendix III (shown in Figure 4). To improve the accuracy for  $Q_u$ , we measure the insertion loss at  $f_o$  (Figure 3). The graph of the relationship between  $Q_u$  and  $Q$  is shown in Figure 4. For example, the bandpass insertion loss of  $f_o$  in Figure 3 (20 dB) when applied to Figure 4 indicates  $Q_u/Q = 1.1$  or

$$4. Q_u = (1.1)(98.26) = 108.$$

The 20 dB insertion loss is an actual measurement and indicates circuit perturbations due to circuit stray coupling capacitance. Stray capacitance affects the calculated inductance value, but not the value of  $Q_u$  because the relationship between insertion loss (IL), filter  $Q$  and inductance (or tank)  $Q_u$  (given by Equation

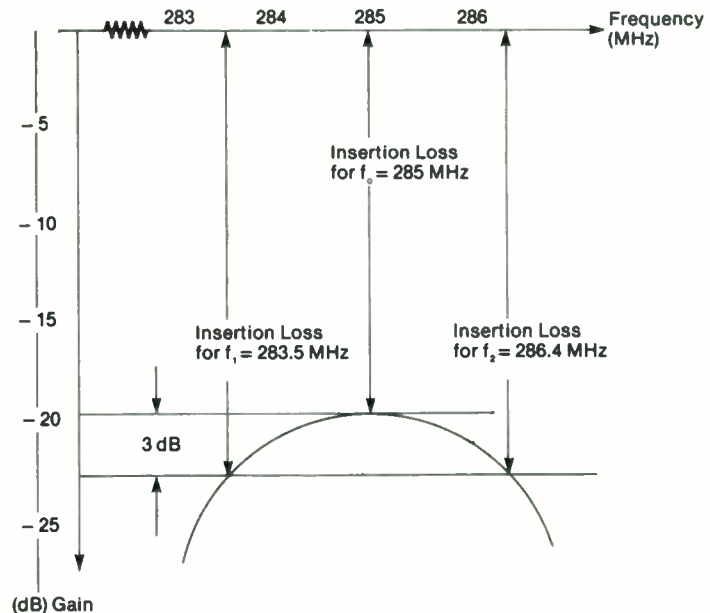


A. The test circuit is a narrowband bandpass filter with accurately measured capacitors.



B. The parallel resonant equivalent of the test circuit.

**Figure 2. The parallel equivalent circuit for the test circuit is used as a basis for calculations.**



**Figure 3. The 3 dB frequencies, which are well defined points, are used to calculate  $f_o$ , and  $Q$ .**

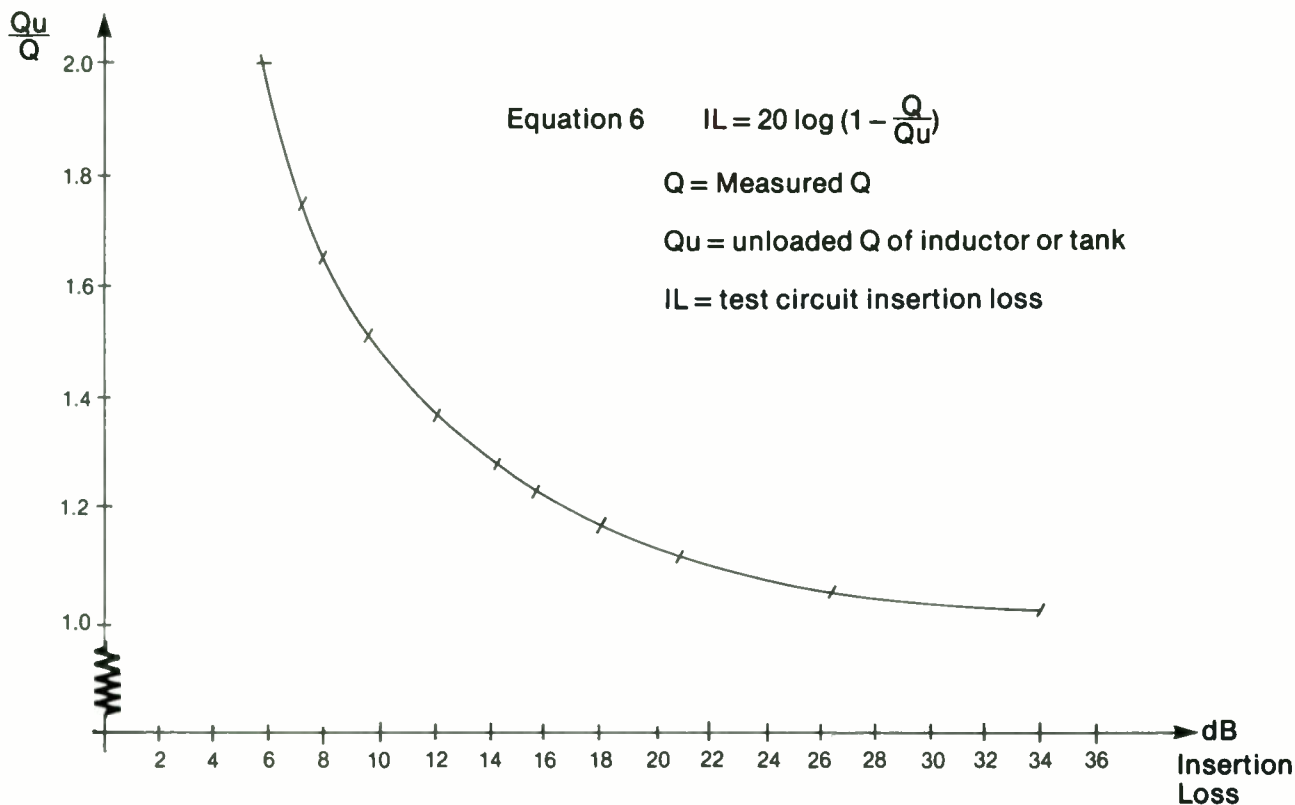
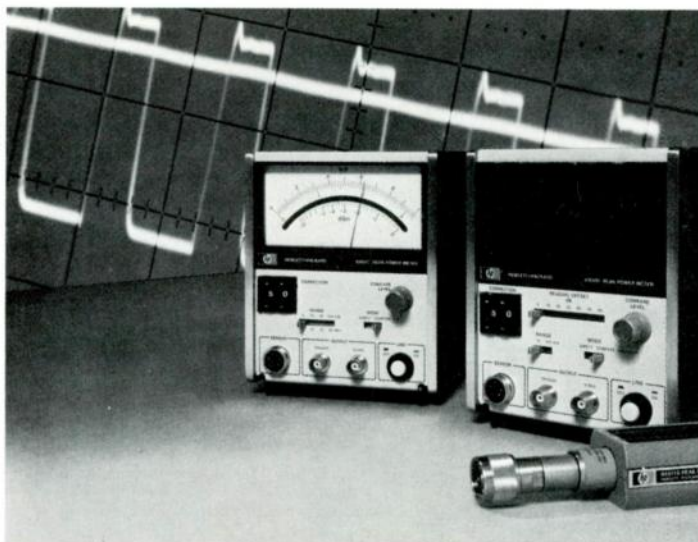


Figure 4. The insertion loss determines the ratio of the unloaded  $Q$  ( $Q_u$ ) to the measured  $Q$ .

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SAS-200/512	200 - 1800 MHz	Log Periodic	SAS-200/560	per MIL-STD-461	Loop - Emission
SAS-200/518	1000 - 18000 MHz	Log Periodic	SAS-200/561	per MIL-STD-461	Loop - Radiating
SAS-200/530	150 - 550 MHz	Broadband Dipole	BCP-200/510	20 Hz - 1 MHz	LF Current Probe
SAS-200/540	20 - 300 MHz	Biconical	BCP-200/511	100 KHz-100 MHz	HF/VHF Crnt. Probe
SAS-200/541	20 - 300 MHz	Bicon'L Collapsible			

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L	$X_L$	$R_{eff}$	$R_L$	$R_C$	$Q_{SP}$	$X_C$	$C_C$	L
	$\omega_0 L$	$QX_L$	$Q_U X_L$	$R_{eff} = \frac{R_L 2R_C}{R_L + 2R_C} \sqrt{\frac{R_C}{50} - 1}$		$\frac{R_C}{Q_{SP}}$	$\frac{1}{X_C \omega_0}$	$\frac{1}{\omega_0 (3.96 + C_U \times 10^{-10}) X_L}$
Nh	ohms	ohms	ohms	ohms		ohms	Pf	Nh
75	134.3	13,196	14,504	292,652	76.5	3825.5	.146	73.33
73.33	131.1	12,903	14,182	284,128	76.498	3714	.150	73.21

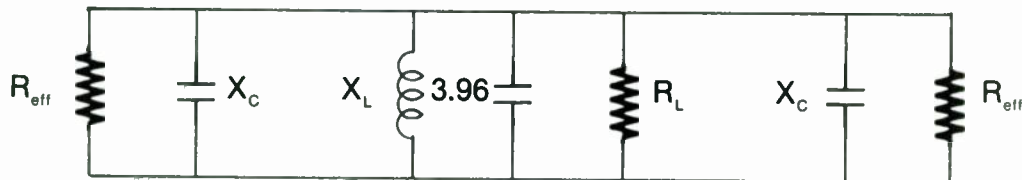


Table 1. The Insertion Loss value is used to resolve the inductance more accurately.



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6) is independent of the coupling capacitors' values or coupling mechanism.

From Figure 4, observe that when the insertion loss is greater than 21dB, the ratio of  $Q_U$  to  $Q$  is less than 1.1, or within ten percent error. The insertion loss value can be used to determine the calculated inductance error due to stray coupling capacitance in the measurement circuit, by using it to solve for  $Q$  and  $Q_U$ , and then solving for the inductance values.

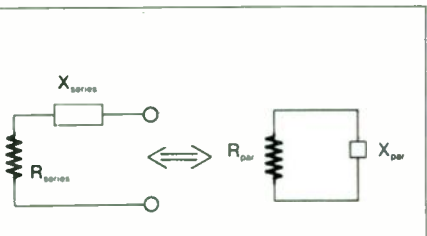
Although a closed form solution for the inductance error is possible, I suggest the following iterative approach (illustrated in Table 1).

1. Use the value of  $L$  to calculate  $X$   
 $X_L = \omega_0 L$
2. Calculate the effective resistance of the circuit  
 $R_{eff} = QX_L$
3. Calculate the inductor resistance  
 $R = Q_U X_L$
4. Calculate the shunt equivalent of the load and source resistances.  
( $R_C$ )

$$R_C = \frac{R_{eff} R_L}{2(R_L - R_{eff})} \text{ from } R_{eff} = \frac{(R)(2R_C)}{2R + R_L}$$

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**Figure 5. The parallel to serial transformation is used to establish the series equivalent circuit.**

5. Calculate the Qsp of the series to parallel transformation

$$Q_{sp} = \sqrt{R_c/50 - 1}$$

from Equation 2

6. Calculate  $X_c = R_c/Q_{sp}$

7. Solve for the coupling capacitance  $C_c$

$$C_c = 1/X_c \omega_0$$

8. Solve for  $L = \frac{1}{2\pi(3.96 + C_c)X_c}$   
 $= 73.2 \text{ uH}$

use the calculated value for L or recycle.

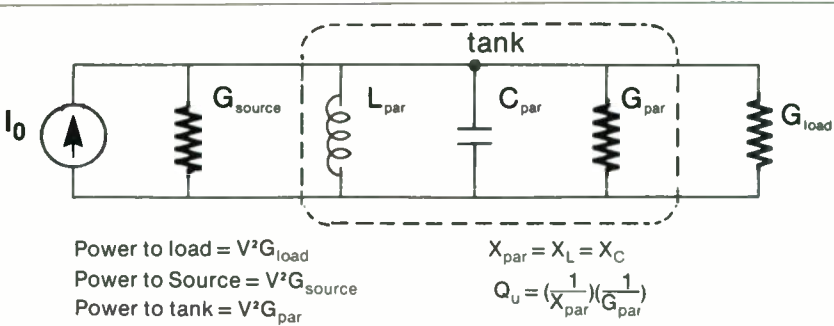
A measurement error in the resonant capacitor also affects the calculated inductance value. For example, an error of 0.1 pF in the capacitance measurement provides less than 1.73 nH uncertainty in the inductance value as shown by Equation 10 (Appendix IV).

$$\Delta L = -1/(\omega_0 C)^2 = -1.72 \text{ nH}$$

where  $\Delta C = 0.1 \text{ Pf}$   
 $\omega_0 = 2\pi \times 285 \times 10^6$   
 $C = (3.96 + 0.15 + 0.15) \text{ pF}$   
 $= 4.26 \text{ pF}$

### Conclusion

A measurement technique to measure inductance and unloaded Q has been illustrated by an example. The measurement is at the frequency of operation and includes stray capacitance effects. The measurements use non-exotic laboratory equipment and relies on an accurate measurement of capacitance, insertion loss and frequency. The basis for the measurement is the construction of a narrow-band filter whose Q approximately equals the unloaded Q of the inductor or tank being measured.



**Figure 6. The Parallel equivalent of the test circuit is a filter whose midband IL relates the Q of the filter to the unload Q ( $Q_u$ ) of the tank.**

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## Appendix I.

Equations for series to Parallel Transformations.

The parallel equivalence of the series load resistance and coupling reactance is shown in Figure 5. The transformation equations are calculated by solving for the series equivalent of the shunt circuit.

$$Z_{\text{series}} = \left( \frac{R_{\text{par}} (jX_{\text{par}})}{R_{\text{par}} + (jX_{\text{par}})} \right) \left( \frac{R_{\text{par}} - jX_{\text{par}}}{R_{\text{par}} - jX_{\text{par}}} \right) \quad \text{Multiplication of Z series to rationalize denominator}$$

$$Z_{\text{series}} = \frac{R_{\text{par}} (X_{\text{par}})^2 + jX_{\text{par}} (R_{\text{par}})^2}{R_{\text{par}}^2 + X_{\text{par}}^2}$$

$$= R_{\text{series}} + jX_{\text{series}} \quad (1)$$

Define Q (of the reactance)

$$Q = \frac{R_{\text{par}}}{X_{\text{par}}} \quad (2)$$

then from Equation 1 and Equation 2

$$R_{\text{series}} = \frac{R_{\text{par}}}{1 + Q^2} \text{ or } R_{\text{par}} = R_{\text{series}} (Q^2 + 1) \quad (3)$$

$$X_{\text{series}} = \frac{X_{\text{par}} Q^2}{Q^2 + 1} \text{ or } X_{\text{par}} = X_{\text{series}} = \frac{(Q^2 + 1)}{Q} \quad (4)$$

for large Q,  $Q \gg 1$

$$X_{\text{par}} = X_{\text{series}}$$

please note from Equations 2, 3, and 4

$$Q = \frac{X_{\text{series}}}{R_{\text{series}}} = \frac{X_{\text{par}} Q^2}{R_{\text{par}}} = \frac{R_{\text{par}}}{X_{\text{par}}} \quad (5)$$

## Appendix II.

The series coupled source (and load) are transformed to their parallel equivalents by applications of the equations derived in Appendix 1.

$$X_{\text{series}} = \frac{1}{2\pi f_0 C} = 5.58 \times 10^3 \text{ ohm} \quad \text{where } f_0 = 285 \text{ MHz}$$

$$C = 0.1 \text{ pF}$$

Applying equation 5

$$Q = \frac{X_{\text{series}}}{R_{\text{series}}} = \frac{5.58 \times 10^3}{50} = 111.7 \quad \text{where } R_{\text{series}} = 50 \text{ ohms}$$

from Equation 3

$$R_{\text{par}} = R_{\text{series}} (Q^2 + 1) = 50 \{ (111.7)^2 + 1 \} = 624,000 \text{ ohms}$$

from Equation 4

$$X_{\text{par}} = X_{\text{series}} \frac{(Q^2 + 1)}{Q^2} = (5.58 \times 10^3) \left( \frac{(111.7)^2 + 1}{(111.7)^2} \right) = 5.58 \times 10^3$$

for large Q, as we've shown

$$X_{\text{par}} = X_{\text{series}}$$

and

$$C_{\text{par}} = C_{\text{series}} = 0.1 \text{ pF}$$

$$C_{\text{resonant}} = \frac{1}{\omega_0^2 L} = \frac{1}{((2\pi(285 \times 10^6))^2 (75 \times 10^{-9}))} = 4.16 \text{ pF}$$

$$C_{\text{resonant}} = C_{\text{par}_{\text{source}}} + C_{\text{par}_{\text{load}}} + C_R$$

## Appendix III.

The insertion loss (IL) of the test circuit relates the filter Q to the  $Q_u$  of the tank, or inductor, by Equation 6.

$$IL = 20 \log (1 - Q/Q_u) \quad (6)$$

where Q = test circuit Q

$Q_u$  = tank or inductor  $Q_u$

The circuit is shown in Figure 6

$P_{\text{actual}}$  = Power delivered to load =  $V^2 G_{\text{load}}$

$$= \frac{(I_0)^2 (G_{\text{load}})}{(G_{\text{source}} + G_{\text{par}} + G_{\text{load}})^2}$$

$$P_{\text{ideal}} = \text{Power delivered to load for } Q_u = \infty \text{ or } G_{\text{par}} = 0 = \frac{(I_0)^2 (G_{\text{load}})}{(G_{\text{source}} + G_{\text{load}})^2}$$

$$\left( \frac{P_{\text{actual}}}{P_{\text{ideal}}} \right)^{1/2} = \frac{G_{\text{source}} + G_{\text{load}}}{G_{\text{source}} + G_{\text{par}} + G_{\text{load}}} \quad (7)$$

from Equation 2

$$Q = \frac{R_{\text{total}}}{X_{\text{par}}} = \left( \frac{1}{X_{\text{par}}} \right) \left( \frac{1}{G_{\text{total}}} \right)$$

or

$$G_{\text{total}} = \left( \frac{1}{Q} \right) \left( \frac{1}{X_{\text{par}}} \right)$$

for  $P_{\text{actual}}$  that is  $0 < G_{\text{par}} < \infty$

$$G_{\text{total}} = G_{\text{source}} + G_{\text{par}} + G_{\text{load}} = 1/QX_{\text{par}} \quad (8)$$

$$G_{\text{total}} = G_{\text{source}} + \frac{1}{Q_u X_{\text{par}}} + G_{\text{load}} = \frac{1}{QX_{\text{par}}}$$

$$\text{since } Q_u = \left( \frac{1}{G_{\text{par}}} \right) \left( \frac{1}{X_{\text{par}}} \right)$$

$$G_{\text{source}} + G_{\text{load}} = \frac{1}{QX_{\text{par}}} - \frac{1}{Q_u X_{\text{par}}} \quad (9)$$

Substitution of Equation 8 and 9 into Equation 7

$$\left( \frac{P_{\text{actual}}}{P_{\text{ideal}}} \right)^{1/2} = \frac{\left( \frac{1}{QX_{\text{par}}} \right) - \left( \frac{1}{Q_u X_{\text{par}}} \right)}{\frac{1}{QX_{\text{par}}}} = 1 - \frac{Q}{Q_u}$$

$$\text{Since Insertion Loss (IL)} = 20_{\log} \left( \frac{P_{\text{actual}}}{P_{\text{ideal}}} \right)^{1/2}$$

$$\text{Then IL} = 20_{\log} \left( 1 - \frac{Q}{Q_u} \right)$$

As the Q of the test circuit approaches the unloaded Q ( $Q_u$ ) of the tank, the insertion loss (IL) increases as shown in Figure 4.

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## Appendix IV.

The inductance error due to capacitance measurement error is given by Equation 10.

$$\Delta L = \left( \frac{-1}{(\omega_0 C)^2} \right) (\Delta C) \quad (10)$$

where  $\Delta C$  = capacitance measurement error

$$\omega_0 = 2\pi f_0 \text{ (285 MHz)}$$

$$C = \text{total shunt capacitance} \\ = (3.96 + .03) \text{ pF} = 4.26 \text{ pF}$$

It is then derived by taking the derivative of the resonance equation.

$$L_C = 1/\omega_0^2 \\ dL \times C + L \times dC = 0 \\ dL = \frac{L}{C} \times dC = -\frac{L}{C^2} \times dC$$

$$dL = \left( \frac{-1}{(\omega_0 C)^2} \right) dC \quad (11)$$

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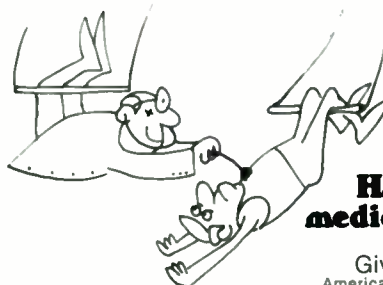
They're afraid the doctor might "find something." This kind of fear can prevent them from discovering cancer in the early stages when it is most often curable.

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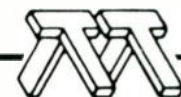
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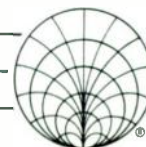
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# Perfecting MOSFET Technology

**B**y perfecting a new MOSFET fabrication technology called Isofet™, Acrian enables power amplifier designers in the VHF and microwave regions to design amplifier stages with higher stage gain, enhanced input/output isolation and better thermal stability.

Though MOSFET technology in the past has been limited to the VHF frequencies — 300 MHz and below, Acrian's Isofet MOSFET devices are suitable for a variety of applications, as high as 4 GHz. As an example, one Acrian Isofet device has demonstrated the capability of delivering 100-watts, pulsed at 1 GHz with a gain of 15 dB.

A comparison with earlier VMOS technology specifications points up clearly the specifications of the Isofet that contribute to its superior performance. Power gain is typically 6 to 10 dB higher. The  $F_T$  is typically four times as high. What's more, the gate-to-drain capacitance ( $C_{RSS}$ ) is typically 1.0 pf — or one-fourth that of other MOSFET devices. This accounts for its superior input/output isolation thereby easing stable amplifier design. There is less likelihood of spurious operation.

Isofet devices are also more rugged,

can survive VSWR of infinity, and are also more tolerant insofar as safe operating region.

The Isofet technology employs a shorter channel, shorter base-width and narrower diffusion. Consequently, since mobility is some 30% higher in the surface region, the  $F_T$  is much higher than in many comparable MOSFET devices — as high as 5 GHz.

Another benefit is the dramatically reduced resistance of the parasitic bipolar transistor in parallel with the FET. It is typically one-sixth as large as the corresponding resistance in a VMOS design. This reduces the likelihood of secondary breakdown.

Isofet MOSFET technology also allows for a higher operating voltage. The lower ON resistance also renders the Isofet less prone to thermally-induced secondary breakdown since the safe operating area of the device is only limited by the breakdown voltage of the device and its thermal dissipation ability.

Acrian's Isofet technology departs markedly from VMOS technology by positioning the gate on the surface of the device, rather than in the "V-groove". Thus the channel is near the surface where mobility is higher, there-

by raising the  $F_T$ . Also, the channel is much shorter, which accounts for the high  $F_T$ .

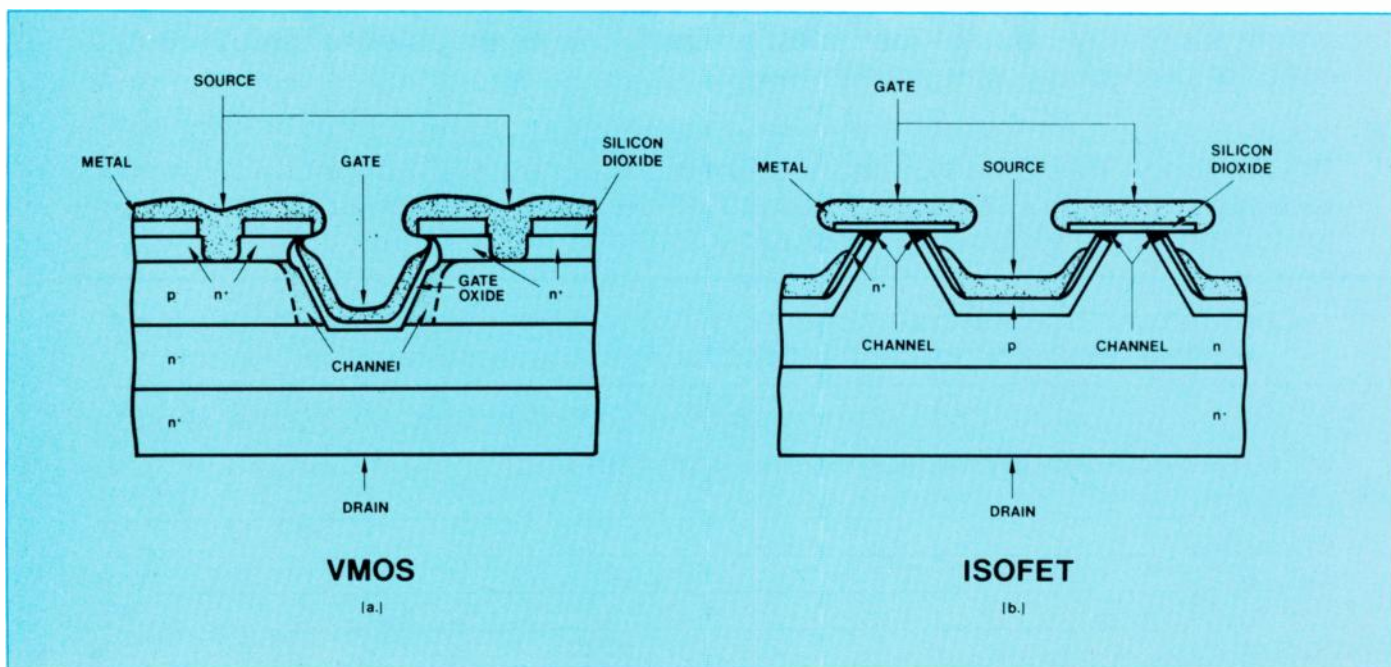
In the early 1970s Westinghouse developed and patented a shadow-isolated VMOS transistor. This approach entailed etching silicon from beneath a silicon-dioxide roof and then using the roof to shadow an evaporated metal that formed the gate. The Westinghouse process entailed elaborate fixturing to aim the evaporated metal properly at the channel under the roof.

This process produced VMOS devices that operated in the 1.0 GHz region. However, this technology was never put into production.

In 1979 Siliconix introduced a VHF shadow-isolated VMOS transistor utilizing a similar oxide roof. However, their device was evaporated in a normal rotating planetary system, completely filling the bottom of the V-groove with metal.

Isofet™, Acrian's shadow DMOS device also utilizes the rotating planetary evaporation technique. However, the Isofet technology differs in that the channel is raised to the surface of the device thereby exploiting the higher surface mobility. What's more, since the channel is parallel to the surface, the channel length is shorter and capacitance is lower.

Isofet is MOSFET technology or metal gate technology. Whereas silicon gate technology is confined to 5.0 MHz, metal gate technology is less lossy than polysilicon. The shadow-isolated overhang acts as a self-aligning mask. Alignment is 'triply' self-aligning. Thus Isofet MOSFETs are cheaper to make since they require fewer masking operations. □





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The UMIL40FT consists of two Isofet devices internally connected in push-pull. It exhibits 10-dB more gain than a bipolar device and 3-dB more gain than MOSFET devices previously available for r-f designs.

The new Acrian Isofet is well suited for AM and FM aircraft/land communication, 2-30 MHz SSB, 30-88 MHz programs, Military AM/FM in the 100-400 MHz region and broadband instrumentation and countermeasures in the 200 to 500 MHz range.

One of the characteristics of the UMIL40FT which appeals to power amplifier designers is that a virtually-flat, power-output profile is readily achieved without the need for gain sloping. This technique has frequently been required in bipolar circuits intended to operate over octave bandwidths. Sophisticated gain sloping networks using R, L and C or quadrature combiners are required with bipolars.

Another feature which simplifies design is the fact that the input impedance can be adjusted to 12.5, 25 or 50 ohms. This eliminates any requirement for complex input matching networks. Simple input transformers are all that are necessary.

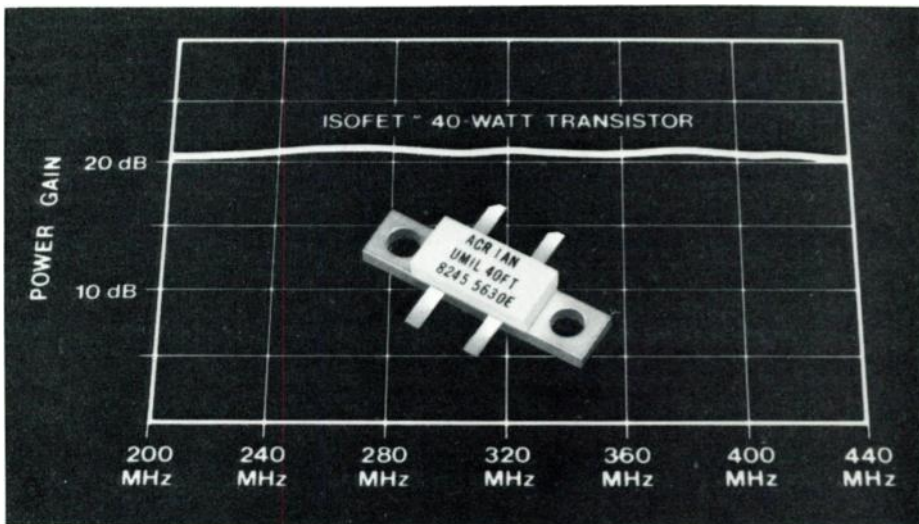
Biasing can be accomplished simply with a resistor. Sensitive tracking diodes with feedback are not required. Also the Acrian UMIL40FT Isofet is more forgiving in regard to VSWR, has a large safe operating area, and has a

maximum dissipation of 100 watts.

FETs have the additional benefit that they can be high-level modulated through simple voltage control of the gate. Thus, simple logic circuits can be utilized to directly modulate the UMIL40FT. There is no need to simul-

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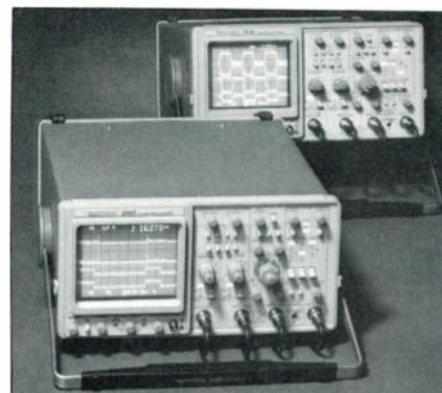
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## 2400 Series Oscilloscopes

Using today's most advanced technologies, Tektronix 2400 Series portable oscilloscopes establish a new industry standard. Two 2400 scopes are being introduced, the 2445 and 2465. The 2445 and 2465, analog scopes with significant benefits for digital applications, are designed to replace the Tektronix 465B and 475. At \$3,140 and \$4,600, the new scopes' broader bandwidths, four channels, extensive CRT readouts, faster sweep speeds and increased timing accuracy set the standard for scopes to come. The 2465 has 300 MHz bandwidth, 500 picosecond per division sweep speed, and an auto-level trigger feature with trigger bandwidth greater than 500 MHz. The 2445 has 150 MHz bandwidth, 1 nanosecond per division

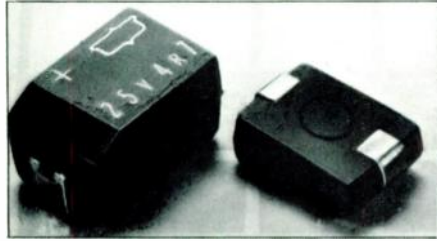
sweep speed, and the same triggering features with trigger bandwidth to 250 MHz. The 20.3-pound scopes have four full-bandwidth vertical channels, each with calibrated scale factors and positioning controls. Another feature



is the channel 1 and 2 propagation delay matching adjustment. **Tektronix, Inc., Beaverton, OR 97077** or **PLEASE CIRCLE INFO/CARD #140.**

### Leadless Aluminum Electrolytic Chip Capacitors

United Chemi-Con, Inc. announces the world's first aluminum electrolytic chip capacitor. The ALCHIP-S is ideal for hybrid circuit and leadless printed circuit board applications. This series is available with a capacitance range of 0.1 to 22  $\mu\text{F} \pm 20\%$ , in six voltages from 6.3 to 50 VDC. Mechanically, the ALCHIP-S utilizes a molded plastic case capable of withstanding 230°C



for 10 seconds, allowing assembly by standard re-flow soldering techniques. Price: \$.07 each in 1,000 quantities. **United Chemi-Con, Inc., Rosemont, IL 60018, INFO/CARD #139.**

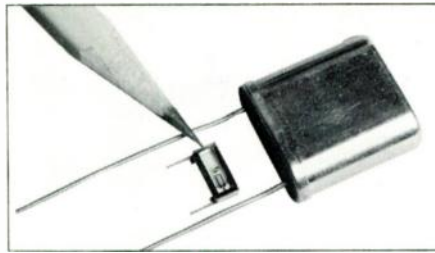
### SPDT Module Switches With Driver

M/A-COM Microwave Circuits, Inc. Hudson, NH announces the MA-8345 series of SPDT module switches with integrated drivers. These hermetically constructed, connectorless units cover the frequency range of .5 to 18 GHz and offer a switching speed of 10 ns. They also offer high isolation and TTL compatibility. The MA-8345 series will find use in RF switching modulation and channel selection. Designed specifically for broadband ECM requirements, their light weight and small size make them an excellent choice for airborne and missile applications. **M/A-COM Microwave Circuits, Inc., Hudson, NH 03051, or INFO/CARD #137.**

### Ultra-Miniature Quartz Crystals

Ultra-miniature quartz crystals that operate at frequencies up to 1.5 MHz have been introduced by Statek Corp. The manufacturer claims that they are the smallest crystals available at any frequency in the range 10 kHz to 1.5 MHz. They are 48 times smaller in volume than the industry-standard HC-33 crystal package and occupy only one-tenth as much board space. This expanded CX-1 Series is available in a hermetically-sealed ceramic package

r.f. design



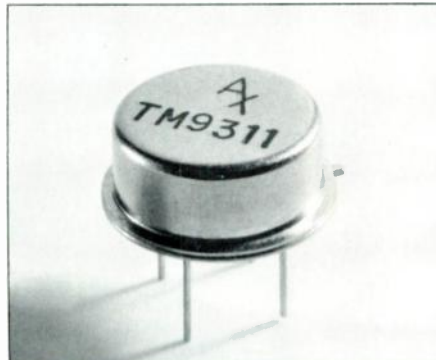
with 0.300-inch lead spacing and in a leadless version for surface mounting. Prices: Less than \$10.00 each in 100 quantity. Delivery: Twenty-three standard frequencies available from stock. **Statek Corp., Orange, CA 92668, or INFO/CARD #138.**

### Plotter Module Adds Graphics To HP-41 Handheld Computers

A low-priced system for generating graphics and bar code with the HP-41 handheld computers is now available from Hewlett-Packard Company. With the introduction of the plotter module at a \$75 list price, HP-41 owners can link their handheld computers to the HP 7470 color graphics plotter and plot charts, graphs and bar code. The HP-41 handheld computer, with the plotter module, is connected to the HP 7470 graphics plotter via the Hewlett-Packard Interface Loop (HP-IL). A new HP-IL-compatible version of the plotter (HP 7470 Option 003) is being introduced concurrently with the plotter module. The new module plugs into the HP-41 and adds a set of commands and functions to the handheld computer that enable the user to plot charts, graphs and bar code. **Call your local Hewlett Packard Sales Office or INFO/CARD #122.**

### RF Amplifier

Amplifonix has announced a new RF Hybrid Amplifier, Model TM9311 in a TO-8 hermetic package. This new design provides a gain of 16 dB typical over the frequency range of 5 to 1000 MHz. Other specifications include: Noise Figure, 2.4 dB typ; Power Out @ 1 dB comp., -1 dBm; Max. VSWR (In/Out); 2:1; Current @ 15V, 9 ma.; Temperature range, -55 to +100°C.



All units meet MIL-STD-883B screening. Price for 1-9 is \$90; delivery from stock. **Amplifonix, Inc., Bristol, PA 19007, or INFO/CARD #135.**

### Micro-Miniature Band Pass Filters

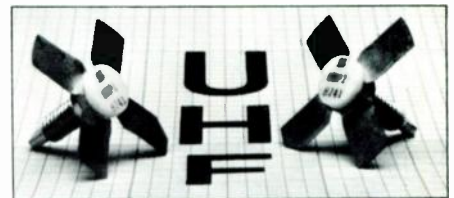
RLC Electronics, Inc. announces the introduction of Micro-Miniature Band Pass Filters. These units offer excellent frequency response characteristics with low insertion loss. The small size capability is provided by utilization of miniaturized High Q devices in a micro-strip mode. Standard units utilize low ripple, Chebyshev design. Other re-



sponses available when desired. Filters are available with spurious free response to 12.4 GHz when required. Prices start at \$250.00 in unit quantity. Delivery of small quantity in 6 weeks. **RLC Electronics, Inc., Mt. Kisco, NY 10549, or INFO/CARD #134.**

### UHF Power MOSFETS

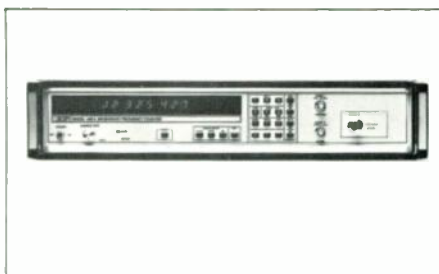
Siliconix has announced the immediate availability of the first commercially available 5 and 10 watt UHF



power MOSFETs. Designated the UMP-1 and UMP-2, the n-channel, enhancement-mode devices can deliver 5 watts and 10 watts, respectively, while operating at frequencies up to 400 MHz. Specified minimum power gains are 10 dB for the UMP-1 and 7 dB for the UMP-2. The introductions of the new UHF parts extends Siliconix MOS-POWER devices beyond the HF/VHF operating frequency range into the UHF region. **Siliconix Inc., Santa Clara, CA 95054, or INFO/CARD #133.**

### Counter For Communications Use

EIP Microwave has announced a new configuration of their 10Hz to 18GHz 545 Microwave Frequency Counter designed especially for the communications industry. The manufacturer claims that this new unit, designated the 545-W29, offers several



advantages to telco transmission engineers and craftsmen in the alignment, repair and verification of microwave communications equipment. These advantages include: 1) An input

YIG filter for frequency selective power measurements. This allows output power to be measured right at the feedhorn test-point for quicker and easier adjustment of transmitter power levels on multiple carrier systems. 2) Measurement of final power and frequency, without turning off modulation. 3) A built-in power meter for direct power readings. 4) Five watts of burn-out protection to prevent counter failure due to accidental overload. Priced at \$7400, the 545-W29 weighs only 20 lbs. EIP Microwave, Inc., San Jose, CA 95134, or please circle INFO/CARD #131.

## Sampling Oscilloscope System

I.T.I. S-500 is a dual trace delayed Equivalent-Time Sampler used with any oscilloscope capable of X-Y display to form a Sampling Oscilloscope system. It provides an input bandwidth of dc-700MHz and a Time Base from 500ps/div. to 100ns/div. A P-700 FET-Probe with dc-500MHz bandwidth can be used to acquire signals from high impedance sources to the sampler.



The S-500 and the P-700 are priced at \$480 and \$63 respectively. International Test Instruments Corp., Cupertino, CA 95015, or please circle INFO/CARD #130.

## Leadless Inductive Components

An expanded line of standard and miniaturized West-Chip™ leadless ceramic chip inductors is offered by West-Cap Arizona, a subsidiary of SFE Technologies, San Fernando, California. They are designed for use in thick film, high-density hybrid microcircuitry, resonant circuits and decoupling applications, as well as in low power signal applications. Mounting is by either solder reflow or conductive epoxy. Core material is ferrite, powdered iron or phenolic, depending upon the inductive value of the device. Standard WCS series inductors, mounted on .125" x .150" ceramic substrate, are supplied in a standard range of .010 uH to 1000 uH. Custom devices are available up to 47 mH. West-Cap Arizona, San Fernando, CA 91341 or INFO/CARD #129.

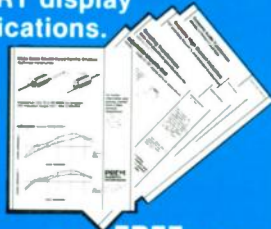
## SMA Coaxial Attenuators

SMA Coaxial Attenuators are now available with a frequency range from D.C. to 26.5 GHz. Precision machining and minimum internal discontinuities eliminate moding problems normally associated with standard SMA connectors in the 18 to 26.5 GHz band. The connectors will mate non-destructively and provide optimum performance with APC-3.5 connectors. Attenuation values from 1 to 30 dB are available in two subminiature sizes. Typical frequency variation is less than  $\pm 0.5$  dB from nominal and VSWR is less than 1.50 up to 26.5 GHz. KDI Pyrofilm Corporation, Whippany, NJ 07981, or INFO/CARD #128.

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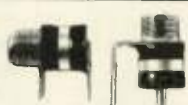
Ferrite power inductors are available from 1.5 to 1500  $\mu$ HY with many configurations.

Prem also manufactures switch mode and linear PC power transformers and filter inductors for powering your particular application. Ask for our standard product listing on these devices.

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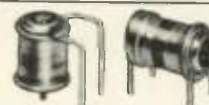
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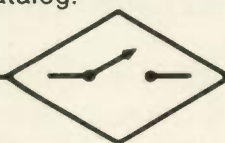
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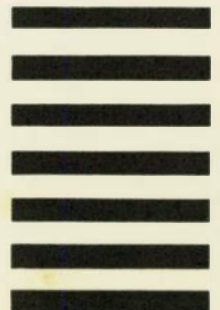
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## RF Plug-Ins HP Sweeper

Hewlett-Packard has expanded the range and performance of its microprocessor-based sweep oscillator family (HP 8350) with the introduction of five new RF plug-ins. The units, which provide new frequency coverages and/or new performance capabilities, are:



### U.S. List Prices and Delivery

Model No.	U.S. List Price	Delivery
HP 83572A (26.5—40 GHz) Option 001-external leveling add	\$14,540 1,605	16 weeks ARO
HP 83592B (0.01—20 GHz)	26,580	12 weeks ARO
HP 83525B (0.01—8.4 GHz)	15,540	4 Weeks ARO
HP 83540B (2.0—8.4 GHz)	10,280	4 Weeks ARO
HP 86251A (7.5—18.6 GHz)	10,780	12 weeks ARO
HP 11869A adapter (for use with with HP 8350)	405	2 weeks ARO
HP 8350A mainframe	4,565	2 weeks ARO

Call local Hewlett-Packard sales office or INFO/CARD #132.

## Microwave Multimeter Active Heads

New Microwave Units are now available for the Microwave Multimeter System.

MODEL	FREQUENCY
7123	1.7—2.3 GHz
7123-S1	2.2—2.7 GHz
7144	6.5—7.2 GHz



With the addition of these Communications and Radar Band Units, the active heads cover all the significant bands in the frequency range from 1.7 to 12.7 GHz. The Narda Microwave Corporation, Hauppauge, N.Y. 11788 or INFO/CARD #126.

## Bench Test Receiver 9 kHz to 30 MHz

The programmable Test Receiver ESH3 is used for measurement and demodulation of Am, DSB, SSB, pulse-modulated and FM signals as well as interfering sinewave and pulse signals in the range 9 kHz to 30 MHz. High overdrive capacity, excellent dynamic range, elaborate measurement/evaluation capabilities and a large number



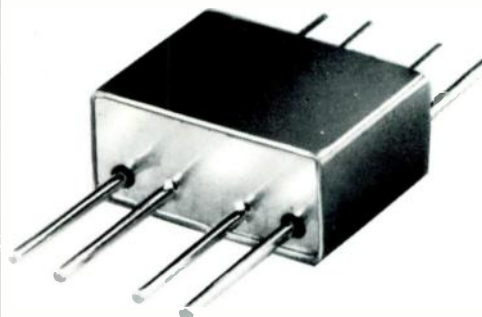
of extras make the ESH3 useful for all applications concerned with radio detection and EMC measurements. The measurement range of -30 to +137 dB (V) makes the ESH3 Test Receiver useful as an automatic, highly accurate, selective voltmeter for laboratories, test departments and inservicing without requiring additional accessories. Polarad Electronics, Inc., Lake Success, N.Y. 11042, or please circle INFO/CARD #127.

## 100 Hz To 1000 MHz Noise Generator

Micronetics has combined the frequency range of 7 different PNG5100 models into one in its new PNG5200 Programmable Noise Generator to cover 10Hz to 1000 MHz frequency band with an output of +10 DBM (10 mw). A computer, operating over the IEEE-488 bus, can control the output level over a range of 99dB in 1dB steps. The source can also be bus-controlled to "on" or "standby" states as is required for system noise-figure measurements. The instrument contains a Solid State Noise Source

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- hi isolation, 40dB
- 1 year guarantee

\*units are not QPL listed

### LMX-113 SPECIFICATIONS

FREQUENCY RANGE, (MHz)  
LO, RF 5-1000  
IF DC-1000

CONVERSION LOSS, dB	TYP	MAX.
one octave from band edge	6.2	7.0
total range	7.0	8.0

ISOLATION, dB	TYP	MIN.
5-50 MHz LO-RF	50	45
LO-IF	45	40
50-500 MHz LO-RF	40	30
LO-IF	35	25
500-1000 MHz LO-RF	30	20
LO-IF	25	17

SIGNAL 1dB Compression Level 0dBm min

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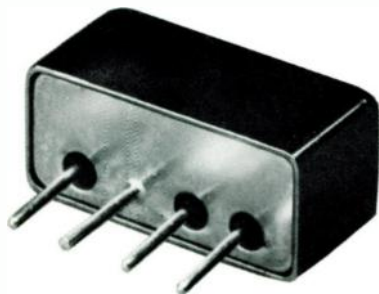
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## TSC-2-1 SPECIFICATIONS

FREQUENCY (MHz)	1-400
INSERTION LOSS, dB	TYP.
(above 3 dB)	
1-10 MHz	0.25
10-200 MHz	0.4
200-400 MHz	0.8
ISOLATION, dB	25
AMPLITUDE UNBAL.	0.2
PHASE UNBAL.	2°
IMPEDANCE	50 ohms

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that produces Gaussian noise with a 5:1 peak-to-RMS ratio, a wideband amplifier, a bandpass filter, a programmable attenuator, and an IEEE-488 bus interface. The requirements of these circuits are stringent, since they must result in the instrument's specified output flatness of  $\pm 2.5$  dB over the frequency range. Micronetics, Inc., Norwood, NJ 07648 or INFO/CARD #125.

## Replacement For Discontinued HP 10811B Crystal Oscillator

Model CO-811B offers mechanical and electrical interface interchangeability with the recently discontinued Hewlett Packard Model HP10811B High Stability 10 MHz Crystal Oscillator. It uses an SC-cut crystal for fast re-stabilization and meets the HP10811B noise floor of  $-160$  dBc/Hz, aging of  $5 \times 10^{-10}$  per day, and dual temperature stabilities of  $2.5 \times 10^{-9}$  total over  $0^\circ\text{C}$  to  $+71^\circ\text{C}$  and  $4.5 \times 10^{-9}$  total over  $-55^\circ\text{C}$  to  $+71^\circ\text{C}$ . A series of less stringent stability models is also offered in the same mechani-



cal package with associated cost savings. Vectron Laboratories, Inc., Norwalk, CT 06850 or INFO/CARD #124.

## "T" Attenuator Module

Alpha Industries, Inc., announces the availability of a new "T" attenuator module (#DAA 5125) that features monolithic array construction with multiple beam lead diode junctions. The device utilizes nitride and oxide passivations for high reliability and is supplied in a variety of resin encapsulated packages, or in chip form, for use in microstrip or stripline appli-

cations. Operating temperature is  $-55^\circ\text{C}$  to  $+150^\circ\text{C}$ , with a power handling of 250 MW. These modules are suitable for switching as well as attenuating applications. Alpha Industries, Inc., Woburn, MA 01801 or please circle INFO/CARD #123.

## RF Wattmeter

A new THRULINE® Directional Wattmeter expands the usual single full-scale power level of its plug-in element to seven overlapping power ranges. Tailored for design or field-



service of CW and FM systems from 200 kilohertz (kHz) to 1000 Megahertz (MHz) and 1/4 watt to 10,000 watts, the new precision instrument uses special elements providing seven levels instead of one covering either 1/3/10/30/100/300/1000 watts or 10/30/100/300/1000/3000/10000 watts with  $\pm 5\%$  accuracy OF READING over a full 37dB power range. The desired range is instantly selectable by a front-panel rotary switch which also includes a battery-level position. Elements are simply rotated for either forward or reflected power measurement. Price of this unique Bird RF Wattmeter is \$495, 7-in-1 Elements are \$125-175. Bird Electronic Corporation, Cleveland (Solon), Ohio 44139, or please circle INFO/CARD #120.

## DC-18 GHz Multi-Position Coaxial Switches

Teledyne Microwave, announces its new line of CS-38 broadband sub-miniature coaxial switches with low power Schottky TTL circuits with decoders. The control input is a 3-bit parallel word that is decoded internally to select one of the 3 to 6 available positions. RF performance at 18 GHz is 1.5:1 VSWR, 0.5 dB insertion loss, and isolation of 60 dB. Options available include indicators, actuator volt-

March/April 1983

ages of 12, 15, 20, and 28 VDC, and power connectors. **Teledyne Microwave, Mountain View, California 94043, or INFO/CARD #118.**

## Application Notes

### RF Hermetic Package Mounting Techniques

An application note on techniques for mounting TRW RF linear amplifiers in hermetic packages is available from TRW Semiconductors. The 4-page note describes guidelines for obtaining a suitable interface between a printed circuit board and either a hermetic single-in-line package or a TO-8 package. High reliability screening capabilities for military applications are also outlined. **TRW Semiconductors, Lawndale, CA 90260, or please circle INFO/CARD #110.**

### Transistor Application Note

California Eastern Laboratories, Inc., offers NEC's Technical Note TN82401 titled "Low Distortion Microwave Linear Power Transistor." This paper described the design and experimental results for high gain and low distortion transistors used in a CATV main cable repeater at up to 500 MHz. Diagrams

are located throughout this informative paper. **California Eastern Laboratories, Inc., Santa Clara, CA 95050, or please circle INFO/CARD #113.**

### OP Amp Application Note

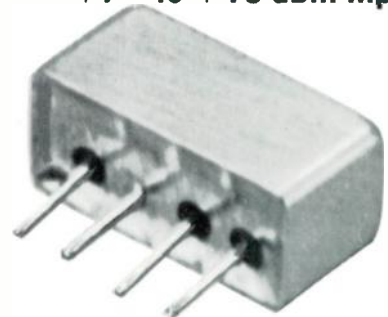
Bill Olschewski, author of many articles on analog circuit design, has written a comprehensive applications note on power op amps covering such subjects as safe operating area, inductive loads, emf generating loads, bridge circuits, speed and position control circuits, cathode ray drive circuits and programmable power supplies. **Apex Microtechnology Corp., Tucson, AZ 85714 or INFO/CARD #112.**

### Auto-Tracking Filter

A new 8 page applications brochure describes Integra's TMF-1800H (2-18.5 GHz) Tunable Microwave Filter. This note illustrates how the Filter can Auto-Frequency Track a signal source or be independently tuned under 488-Bus Control. With this filter, the harmonic output level of Sweepers or Synthesizers can be reduced to a -55 or -70 dBc level. The Tunable Filter can also Offset frequency track a signal source, making it extremely useful for test system applications. **Integra Microwave, Santa Clara, CA 95051, or INFO/CARD #111.**

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+1 to +15 dBm input



**1 to 1000 MHz**  
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- low conversion loss, 13 dB
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- **ruggedly constructed MIL-M-28837 performance\***

\*Units are not QPL listed

### SK-2 SPECIFICATIONS

#### FREQUENCY RANGE, (MHz)

INPUT 1-500

OUTPUT 2-1000

#### CONVERSION LOSS, dB

1-100 MHz TYP. 13 MAX. 15

100-300 MHz TYP. 13.5 MAX. 15.5

300-500 MHz TYP. 14.0 MAX. 16.5

#### Spurious Harmonic Output, dB

2-200 MHz F1 TYP. -40 MIN. -30

F3 -50 -40

200-600 MHz F1 -25 -20

F3 -40 -30

600-1000 MHz F1 -20 -15

F3 -30 -25

For complete specifications and performance curves refer to the 1980-1981 Microwaves Product Data Directory, the Goldbook or EEM.

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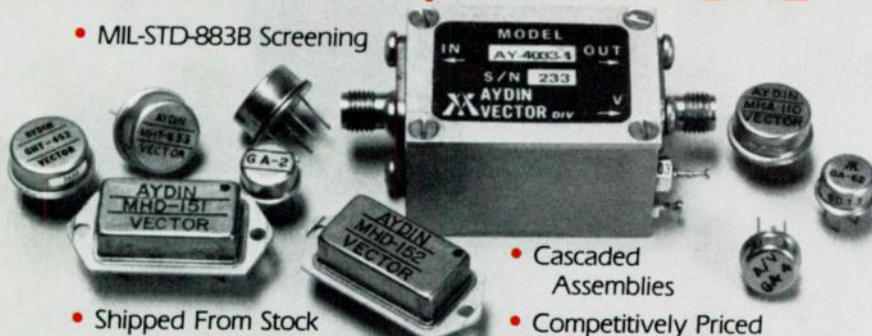
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78-3 REV A

INFO/CARD 64

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Series	Freq Range (MHz)	Typ. Gain (dB)	Noise Figure Range (dBm)	Power Output Range (dBm)	VSWR		Input Power		Package Style
					In	Out	Vdc	Ima	
<b>GA</b> Single-Stage	kHz-400	13	4.5-6.0	5-15	2.0	2.0	+15 typ.	17-70 (range)	TO-12
<b>MHT</b> Single-or Multi-Stage	5-1000	14	2.5-7.0	-2 to +23	2.0	2.0	+15 typ.	10-105 (range)	TO-8
<b>GHT</b> Single-or Multi-Stage	5-400	13	4.0-7.0	5-15	2.0	2.0	+15 typ.	17-70 (range)	TO-8
<b>MHD</b> Multi-Stage	1-500	21	2.5-6.5	5-20	1.6 (typ.)	1.6 (typ.)	+15 typ.	50-115 (range)	4-pin DIP

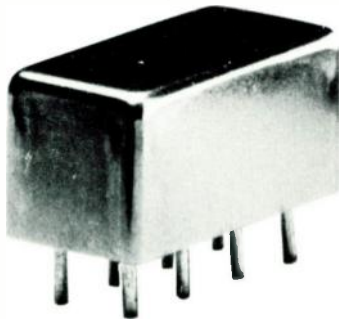
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- miniature 0.4 x 0.8 x 0.4 in.
- **MIL-M-28837/1A performance\***
- low conversion loss 6.0 dB
- high isolation 25 dB

## SRA-220 SPECIFICATIONS

FREQUENCY RANGE, (MHz)			
LO, RF	.05 - 2000		
IF	.05 - 500		
CONVERSION LOSS, dB			
One octave from band edge	TYP. 6.0	MAX. 7.5	
Total range	TYP. 7.0	MAX. 9.0	
ISOLATION, dB			
.05-.5	LO-RF	TYP. 25	MIN. 20
	LO-IF	25	20
.5-1000	LO-RF	40	30
	LO-IF	40	30
1000-2000	LO-RF	30	20
	LO-IF	25	15

Signal 1 dB Compression level +3dBm

For complete specifications and performance curves refer to the 1980-1981 Microwaves Product Data Directory, the Goldbook or EEM.

\*units are not QPL listed

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## New Literature

### Attenuators Brochure

Weinschel Engineering has published a new document entitled EVERYTHING YOU WANTED TO KNOW ABOUT MIL-A-3933D ATTENUATORS BUT WERE AFRAID TO ASK. This informative synopsis presents all the aspects of specifying attenuators in accordance with MIL-A-3933. It answers such questions as, "What is the impact to the user?", defining CLASS I, II, III and IV. This 10 page document details the requirements invoked when adherence to MIL-A-3933 is requested including: First Article Inspections, Screening Inspections, Group A Inspection, Group B Inspection, Procurement Examples and Test Flow Charts for CLASS I, II, III and IV Attenuator Inspection. As an added extra, a complimentary copy of MIL-A-3933D with all current slash sheets will accompany this informative document. **Weinschel Engineering, Gaithersburg, MD 20877 or please circle INFO/CARD #109.**

### Product Catalog

Kay Elemetrics Corp. new 20 page catalog describes its full line of attenuators, switches and R.F. Test

equipment. The Attenuator product line includes Miniature Step, Standard Step, Rotary, Continuously Variable, Audio and OEM Programmable attenuators. Also described are the R.F. Mega switches, Pulsed Carrier Generators and Sweep Frequency Generator. **Kay Elemetrics Corp., Pine Brook, N.J. 07058, or INFO/CARD #108.**

### Communications Antenna Catalog

A new catalog of commercial, industrial and military HF, VHF and UHF communications antennas and antenna systems is available from Hy-Gain, a division of Telex Communications, Inc. Specifications are stated in both the English and metric systems for antennas ranging from tiny quarter-wave whips for mobile applications to giant arrays requiring multiple acre installation for worldwide communications. When applicable, the U.S. military nomenclature and national stock numbers are included. **Hy-Gain, Lincoln, NE 68505 or please circle INFO/CARD #107.**

### SIPMOS Catalog

The new 1982/83 SIPMOS Small Signal Transistor Catalog has been released by Siemens Components, Inc. The 30-page guide contains sec-

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INFO/CARD 66

tions on SIPMOS technology, maximum ratings, characteristics, bi-polar comparisons, data sheets on the eight configurations and an interpretation of technical data. The transistors presented in the catalog are Siemens' initial offering of small signal SIPMOS FETs. Low drive requirements and fast switching speeds make them ideally suited to interface directly with logic circuits and microprocessors for control of motors, switching power supplies and drives for high powered transistors. **Siemens Components, Inc., Iselin, NJ 08830, or INFO/CARD #106.**

### RF Coaxial Connector Catalog

AVA's new 82-5 catalog features F, UHF, BNC and Twinax connectors and the new Computer Switches. Crimp-type connectors are highlighted in the 16 page, two-color catalog. UG type connectors are cross-referenced in the handy index. Adapters for a variety of connector types are also shown in the revolutionary-for-AVA catalog. **AVA Electronics Corp., Drexel Hill, PA 19026 or INFO/CARD #105.**

### Cable Product Catalog

A new catalog giving complete data on local area network cables, computer cables, and fiber optic cables is now available from the RF Cable Products Division of Times Fiber Communications, Inc., an affiliate of Insilco Corporation. The 24-page illustrated catalog features a four page selection guide and application notes on coaxial cables which engineers and purchasing agents will find most helpful. **Times Fiber Communications, Inc., Wallingford, CT 06492, or please circle INFO/CARD #104.**

### YIG Devices Catalog

A new 48-page YIG Devices catalog is available from Watkins-Johnson Company. The catalog includes the company's full line of YIG-tuned oscillators, filters, integrated devices and harmonic generators. In addition to product listings, the catalog offers a technical introduction to YIG devices, selection guides and application notes. **Watkins-Johnson Company, Palo Alto, CA 94304, or INFO/CARD #103.**

### Solid State Products Catalog

An eight page catalog describing Addington solid state products is available from Eaton Corporation's Communications Products Division. Among the products described are the Addington line of high reliability, thin film VCO's, mixer up/down converter subsystems, and RF cable and connector products. In addition, the brochure describes the division's ad-

vanced capabilities, equipment and quality assurance programs. **Eaton Corp., Sunnyvale, CA 94086, or please circle INFO/CARD #102.**

### Video Delay Lines & Filters Catalog

Allen Avionics has issued a new 8 page catalog, No. 16V, encompassing the firm's complete line of Video and Pulse Delay Lines & Video Filters. Introduced is Hum Eliminator HEC1000, which eliminates hum and other interference in video lines caused by differences in ground potential. Also presented are Random Noise Measurement Networks and Pre-emphasis and De-Emphasis Wave Shaping Networks. A new Delay Equalized Lowpass Filter VSL4P5, is also introduced. **Allen Avionics, Inc., Mineola, NY 11501, or INFO/CARD #101.**

### Thermistor Line Catalog

Designers planning to use negative temperature coefficient thermistors will find a wealth of useful information condensed in a new "pocket-size" catalog available from Dale Electronics. The "catalog" details the products of the Dale Western Thermistor Division. Specifications are given for most frequently specified styles, many of which are available for applications where interchangeability is required. Drawings are provided showing "typical" sensor assemblies which are available from Dale. Included in the "pocket catalog" are RT conversion tables and background information on thermistor applications. **Dale Electronics, Inc., Columbus, NE 68601, or INFO/CARD #100.**

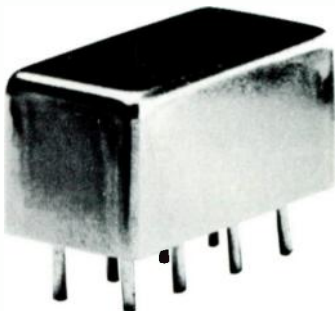
### Test Instruments Brochure

A new brochure describing telecommunications test instruments for digital radio, fiber optics, satellite communications systems and digital modems is available from Scientific-Atlanta, Inc. **Scientific-Atlanta, Inc., Atlanta, GA 30348, or INFO/CARD #99.**

### Capacitor Catalog

A recently revised catalog with all of the newest capacitor offerings from TRW Capacitors is now available. This catalog contains all the information needed to select the appropriate type of capacitor for typical applications. Included are sections on the following types of wound capacitors: metallized polycarbonate dielectric, metallized polypropylene, metallized polyester, polyester dielectric, polystyrene dielectric, and polyester dielectric RC network. **TRW Capacitors, Ogallala, NE 69153, or please circle INFO/CARD #97.** ☐

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only \$11<sup>95</sup> (5-49)**

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- low insertion loss, 0.85dB
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- flat coupling,  $\pm 0.5$ dB
- miniature, 0.4 x 0.8 x 0.4 in.
- hermetically-sealed
- 1 year guarantee

\*Units are not QPL listed

### PDC 10-1 SPECIFICATIONS

FREQUENCY (MHz) 0.5-500			
COUPLING, dB 11.5			
INSERTION LOSS, dB	TYP.	MAX.	
one octave band edge	0.65	1.0	
total range	0.85	1.3	
DIRECTIVITY, dB	TYP.	MIN.	
low range	32	25	
mid range	32	25	
upper range	22	15	
IMPEDANCE	50 ohms.		

For complete specifications and performance curves refer to the Microwaves Product Data Director, the Goldbook, EEM, or Mini-Circuits catalog

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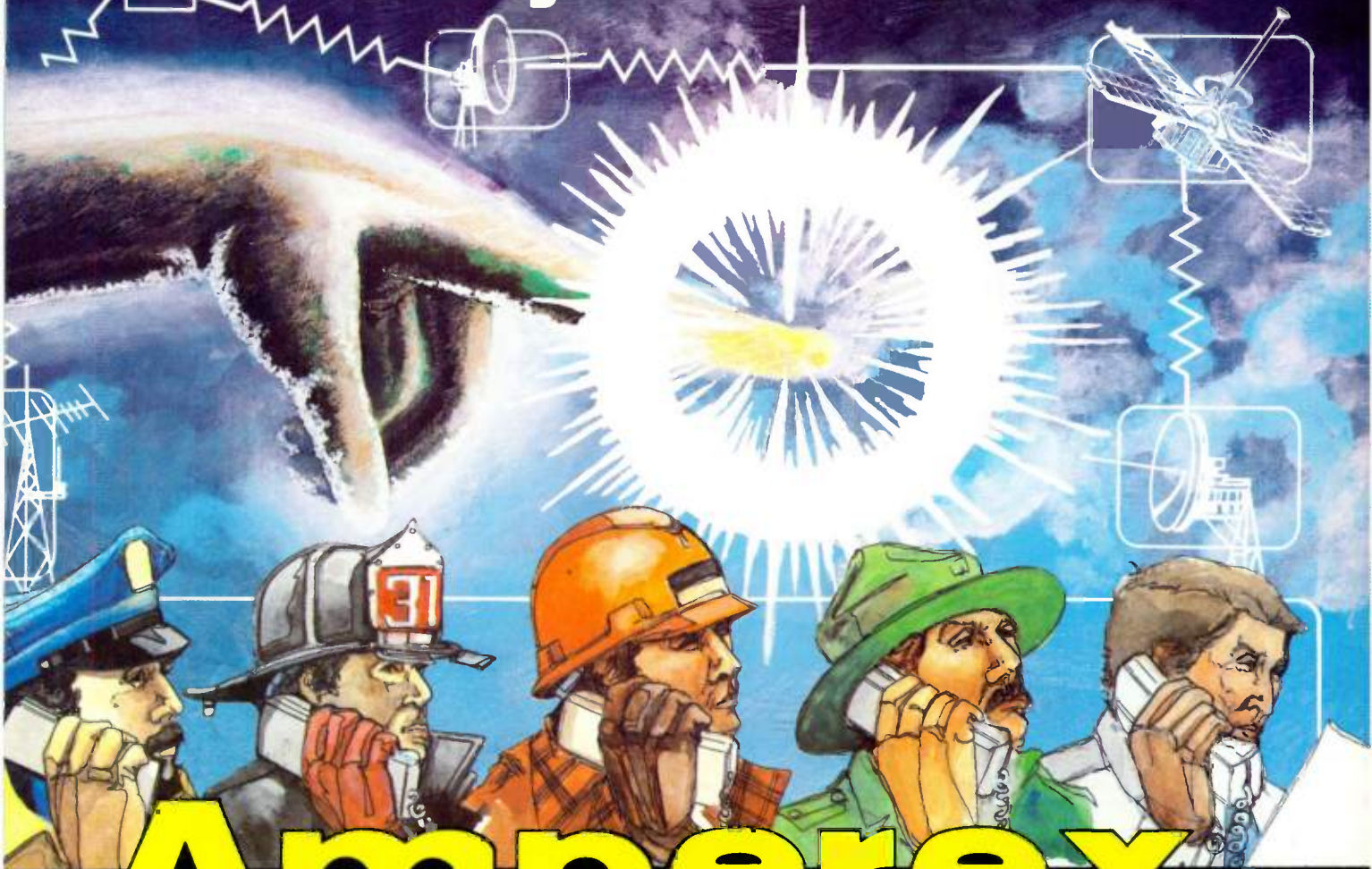
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**APPLICATION:** 900 MHz  
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**POWER:** 0.5 Watts  
**GAIN:**  $\geq 9.0$  dB  
**FREQUENCY:** 900 MHz  
**V<sub>CE</sub>:** 12.5 Volts  
**EFFICIENCY:**  $\geq 50\%$

INFO/CARD 1

• BLV 91

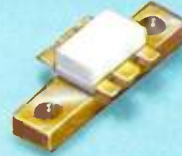


SOT 172 900 MHz

**APPLICATION:** 900 MHz  
 Radio Telephone  
**POWER:** 2 Watts  
**GAIN:**  $\geq 6.5$  dB  
**FREQUENCY:** 900 MHz  
**V<sub>CE</sub>:** 12.5 Volts  
**EFFICIENCY:**  $\geq 50\%$

INFO/CARD 2

• BLV 93



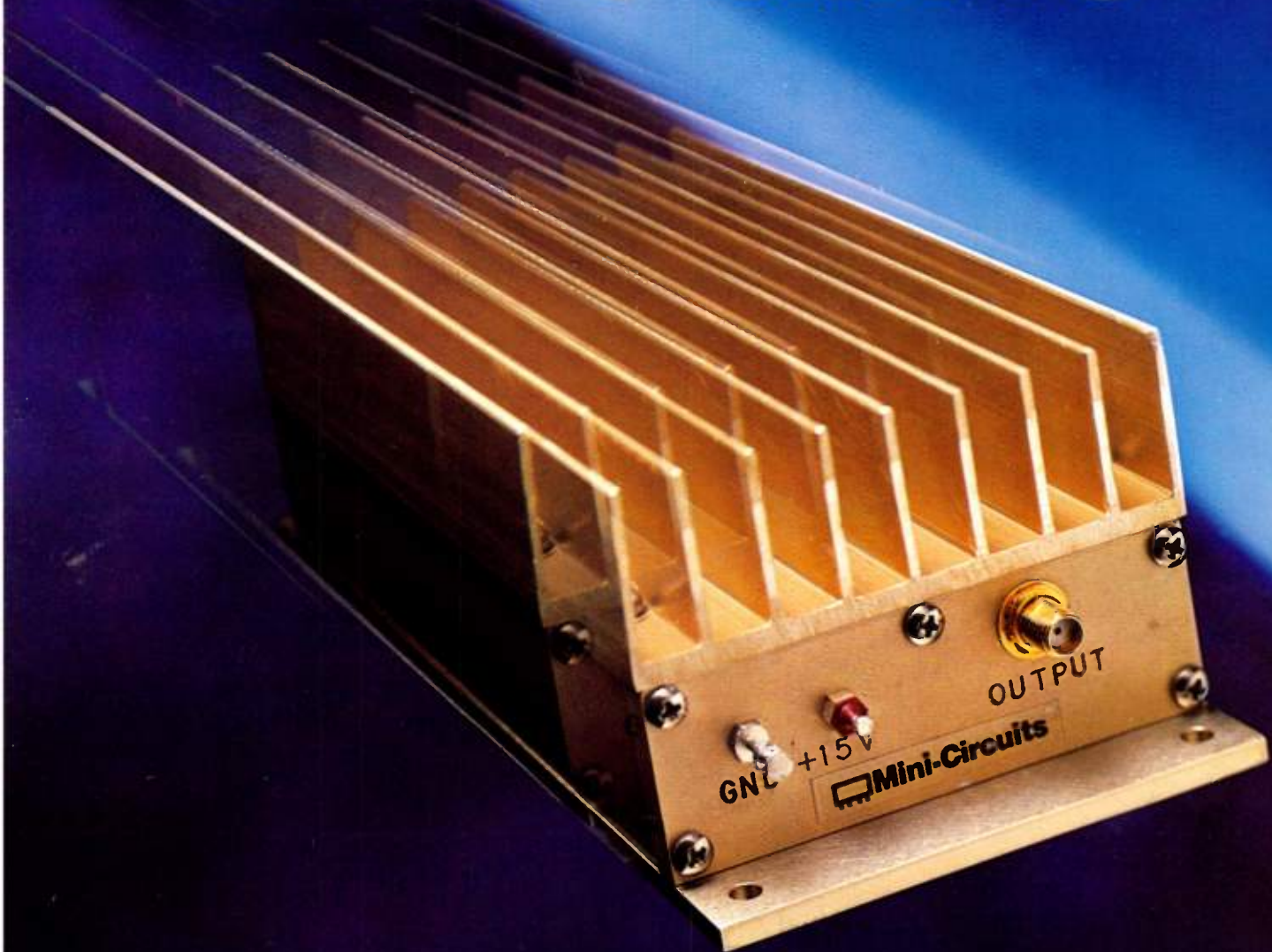
SOT 171 900 MHz

**APPLICATION:** 900 MHz  
 Radio Telephone  
**POWER:** 8 Watts  
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**V<sub>CE</sub>:** 12.5 Volts  
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### ZHL-42 SPECIFICATIONS

Frequency.....	.7 to 4.2 GHz
Gain.....	30dB Min.
Gain Flatness.....	$\pm 1.0$ dB
Power Out @ 1dB CP.....	+29dBm Min.
VSWR In/Out.....	2.0:1 Max.
Noise Figure.....	7.5dB
Supply.....	+15V @ 690mA
Third Order Intercept.....	38 dBm Min.
Second Order Intercept.....	48 dBm Min.
Size.....	7" x 3 1/4" x 2 1/8" h

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